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This journal records developments in the application of electronics in science, industry, and the home. It contains information on the properties and performance of solid-state devices, valves and tubes, and associated components and assemblies. The practical circuits given are designed to provide the best working conditions and performance for the devices used. The design principles may be used (so far as device ratings allow) for adaptation to meet particular requirements. Current data should be consulted before any circuit is considered for production. Mullard Limited can normally supply only those devices or assemblies listed in the Company's current catalogues. Availability in quantity should be confirmed.

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Transistor line timebase for 110° monochrome receivers

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UDC 621.397.332.12 [612.382.3]

The output stage operates from a 150V stabilised or unstabilised h.t. supply, and overscans a 20in or 24in 3:4 aspect ratio picture tube for 625-line receivers by 7%. The e.h.t. supply of 18kV is developed by a high voltage rectifier diode. The circuit employs one high voltage silicon transistor in the output stage, without a separate efficiency diode. Scan rectification is performed by a fast recovery silicon rectifier diode and up to 20W of l.t. power is available for other circuits in the receiver. A new design driver transformer having a high leakage inductance is employed in the driver stage to eliminate a separate choke in the output transistor base drive circuit.

INTRODUCTION

A description is given of a transistor line timebase for monochrome receivers that overscans by 7% a 20in or 24in 3:4 aspect ratio picture tube at an e.h.t. of 18kV. The timebase, shown in Fig. 1, operates with the output stage stabilised supply of 150V maintained by a shunt stabiliser circuit (Ref. 1), and with associated stages operating from the 250V unstabilised h.t. supply. Alternatively, the timebase functions with a thyristor stabilised 150V supply. Component changes for this, and for unstabilised operation of the output stage, are indicated in the components list.

The timebase is described in three main sections: line oscillator stage, line driver stage, and line output stage. A fourth section on flashover protection is also included.

LINE OSCILLATOR STAGE

The oscillator stage is basically the same as that described in 'Transistor 90° colour line deflection and e.h.t. circuit' (Ref. 2). A sinewave oscillator of the LC type is chosen for the line oscillator, in which its frequency and phase are controlled by a variable reactance stage. Both good frequency stability and an output waveform of constant mark/space ratio are obtained. The latter is necessary to maintain the correct operation of the line output stage.

The oscillator is similar to the conventional Hartley circuit in that it has a tapped inductor between the collector and base networks. Capacitor C_8 is connected between an effective a.c. earth and the base of TR_2 and forms a capacitive divider network with C_{10}. This network develops a suitable waveform at the base of TR_2.

A transistor TR_1 is connected across the resonant network and functions as an inductive reactance stage (Ref. 2) controlling the frequency and phase of the oscillator. The phase reference sawtooth waveform for the flywheel circuit (Ref. 2) is obtained from a negative line flyback pulse from the line transformer; the flywheel circuit is controlled by a d.c. output from the TBA550 integrated circuit.

The TBA550 signal-processing integrated circuit used in the flywheel circuit, incorporates a bipolar switch as a phase detector and frequency discriminator (Ref. 2). However, a conventional double-diode phase detector (Fig. 1) may be employed instead of the TBA550 without changing the circuit operation or component values.

P-N-P transistors BC158 are used for TR_1 and TR_2 so that direct coupling to the driver transistor TR_3 can be used. Fig. 2 shows the V_{CE} waveform for the reactance transistor TR_1, and Fig. 3 the V_{BE} and V_{CE} waveforms of the oscillator transistor TR_2.

LINE DRIVER STAGE

The driver stage is designed for the 250V unstabilised h.t. supply. This supply is taken in preference to the line output 150V h.t. supply, derived from the 250V by a dropping resistor, because any open-circuit fault in the output stage would increase the driver stage h.t. from 150V to 250V and damage the transistor TR_3.

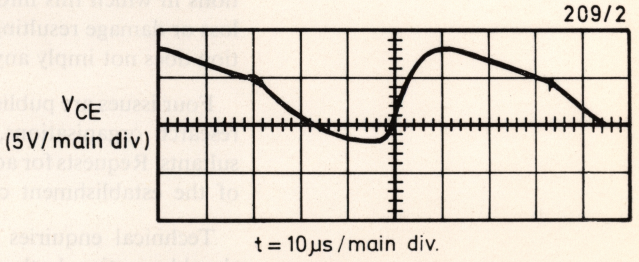

Fig. 2—Reactance transistor TR_1 (BC158) waveform

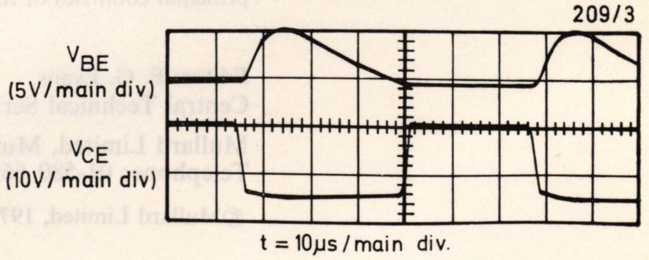

Fig. 3—Oscillator transistor TR_2 (BC158) waveforms

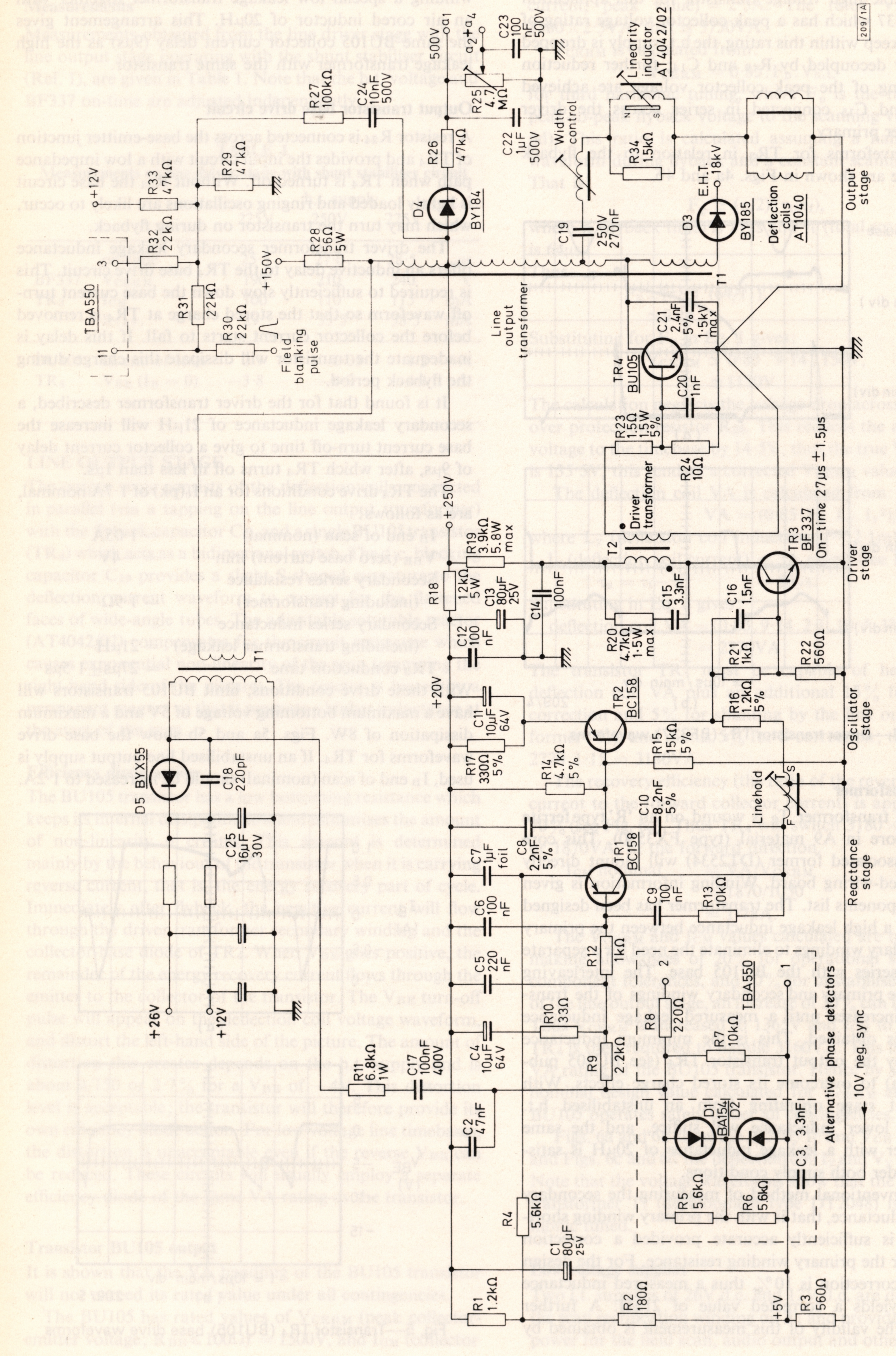

Fig. 1—Transistor line timebase for monochrome receivers

A suitable high voltage transistor for this application is the BF337 which has a peak collector voltage rating of 250V. To keep within this rating, the h.t. supply is dropped and partly decoupled by R_{19} and C_{14}; further reduction and shaping of the peak collector voltage are achieved by R_{20} and C_{15} connected in series across the driver transformer primary.

The waveforms for TR_3 in relation to the flyback (F/b) pulse are shown in Figs. 4a and 4b.

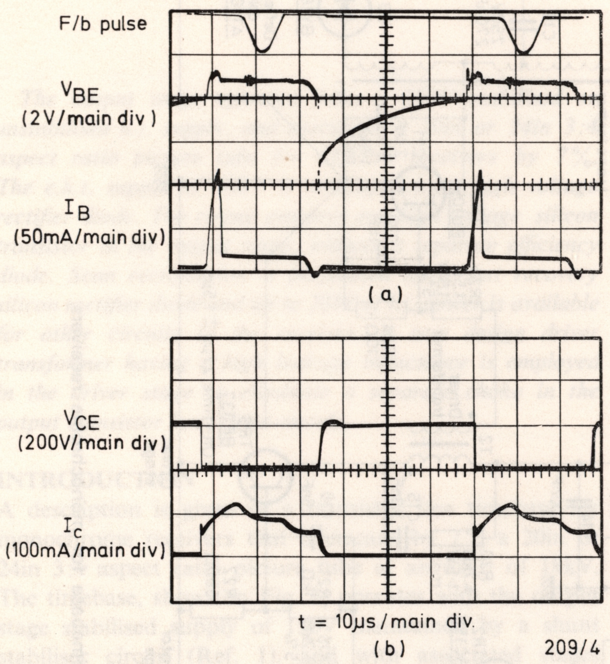

Fig. 4—Driver transistor TR_3 (BF337) waveforms

Driver transformer

The driver transformer T_2 is wound on an 'R' type ferrite inductor core in A9 material (type FX3550). This core with its associated former (DT2534) will mount directly on a printed-wiring board. Winding information is given in the components list. The transformer has been designed to provide a high leakage inductance between the primary and secondary windings to eliminate the need for a separate choke in series with the BU105 base. The interleaving between the primary and secondary windings of the transformer is increased until a measured leakage inductance of 20μH is obtained. This is the minimum inductance required by the output transistor TR_4 (see BU105 published data) to overcome its stored charge effects. With the output stage operating from an unstabilised h.t. supply, a lower inductance will suffice, and the same transformer with a leakage inductance of 20μH is satisfactory under both supply conditions.

The conventional method of measuring the secondary leakage inductance, that is with the primary winding short-circuited, is sufficiently accurate provided a correction is made for the primary winding resistance. For the design given the correction is 10%, thus a measured inductance of 23μH yields a corrected value of 21μH. A further check on the validity of this measurement is obtained by winding a special low leakage transformer together with an air cored inductor of 20μH. This arrangement gives the same BU105 collector current delay (9μs) as the high leakage transformer with the same transistor.

Output transistor base drive circuit

A resistor R_{24} is connected across the base-emitter junction of TR_4 and provides the input circuit with a low impedance path when TR_4 is turned off. Without R_{24} the base circuit is lightly loaded and ringing oscillations are likely to occur, which may turn the transistor on during flyback.

The driver transformer secondary leakage inductance offers an inductive delay to the TR_4 base drive circuit. This is required to sufficiently slow down the base current turn-off waveform so that the stored charge at TR_4 is removed before the collector current starts to fall. If this delay is inadequate the transistor will dissipate this charge during the flyback period.

It is found that for the driver transformer described, a secondary leakage inductance of 21μH will increase the base current turn-off time to give a collector current delay of 9μs, after which TR_4 turns off in less than 1μs.

The TR_4 drive conditions for an I_c(pk) of 1·7A nominal, are as follows:

I_B end of scan (nominal)	= 1·05A
V_{BE} (zero base current) min	= −4V
Secondary series resistance (including transformer)	= 1·9Ω
Secondary series inductance (including transformer leakage)	= 21μH
TR_3 conduction time	= 27μs ± 1·5μs

With these drive conditions, limit BU105 transistors will have a maximum bottoming voltage of 5V and a maximum dissipation of 8W. Figs. 5a and 5b show the base drive waveforms for TR_4. If an unstabilised line output supply is used, I_B end of scan (nominal) should be increased to 1·2A.

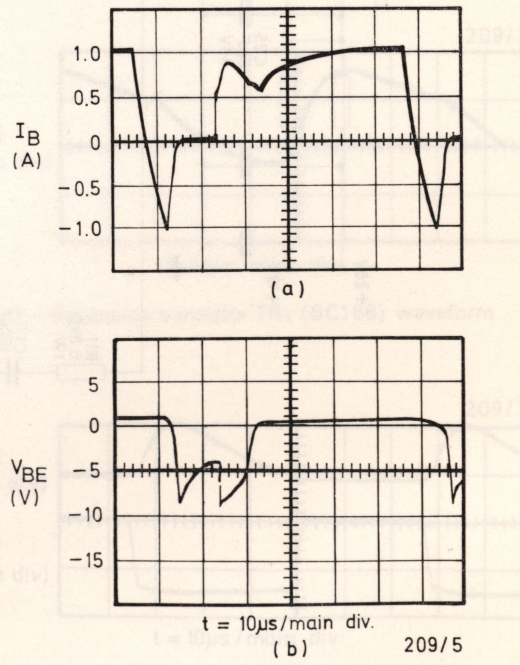

Fig. 5—Transistor TR_4 (BU105) base drive waveforms

Measurements

Measurements obtained from the line driver stage with the line output stage operating with the shunt stabiliser circuit (Ref. 1), are given in Table 1. Note that the h.t. voltage and BF337 on-time are adjusted independently.

TABLE 1

Measurements on line driver stage with shunt stabiliser circuit

		h.t. supply			
		225V	250V	275V	
	on-time	25·5	27·0	28·5	μs
BF337	V_{CERM}	190	210	240	V
TR_3	I_{CM}	100	110	115	mA
	$I_{C(AV)}$	29	33·5	38·5	mA
BU105	I_B end of scan	0·9	1·05	1·2	A
TR_4	V_{BE} ($I_B = 0$)	−3·8	−4·0	−4·1	V

LINE OUTPUT STAGE

The output stage consists of the deflection coils connected in parallel (via a tapping on the line output transformer) with the flyback capacitor C_{21} and a single BU105 transistor (TR_4) which acts as a bidirectional switch. The d.c. blocking capacitor C_{19} provides a slight S-shaped curvature to the deflection current waveform to correct for the flattened faces of wide-angle tubes. The adjustable saturable reactor (AT4042/02) compensates for the circuit resistance which causes exponential non-linearity of the scan (cramping the right-hand side of the picture). The reactor is biased by a permanent magnet so that it presents a higher inductance at the start of the scan.

Efficiency diode action

The BU105 transistor has a low bottoming resistance which keeps its internal dissipation low and minimises the amount of non-linearity it creates. This amount is determined mainly by the behaviour of the transistor when it is carrying reverse current, that is, the energy recovery part of cycle. Immediately after flyback, the negative current will flow through the driver transformer secondary winding and the collector-base diode of TR_4. When V_{BE} goes positive, the remainder of the energy recovery current flows through the emitter to the collector of the transistor. The V_{BE} turn-off pulse will appear on the deflection coil voltage waveform, and distort the left-hand side of the picture. The amount of distortion this creates depends on the h.t. supply and is about 4/150 or 2·7% for a V_{BE} of −4V. This distortion level is acceptable; the transistor will therefore provide its own efficiency diode action. For low voltage line timebases, the distortion is unacceptable even if the reverse V_{BE} can be reduced. These circuits will usually employ a separate efficiency diode of the same VA rating as the transistor.

Transistor BU105 output

It is shown that the VA handling of the BU105 transistor will not exceed its rated value under all contingencies.

The BU105 has rated values of V_{CERM} (peak collector-emitter voltage; $R_{BE} \leqslant 100\Omega$) = 1500V, and I_{CM} (collector current, peak value) = 2·5A. The rated VA is then 1500 × 2·5VA, that is 3750VA.

V_{CERM} is calculated from:

$$V_{CERM} = 0{\cdot}85 \,.\, F_p \,.\, V_{h.t.} \qquad \ldots(1)$$

for third harmonic tuning, where F_p is the ratio of the peak-to-peak flyback voltage to the scanning voltage (Fig. 6d). This ratio is calculated assuming a half-sinusoidal waveform during flyback and a constant scanning voltage. That is:

$$F_p \simeq (\pi/2)\,(\tau_t/\tau_f), \qquad \ldots(2)$$

where τ_f (flyback time) is 11μs and τ_t (total scanning time) is 64μs.

Thus:

$$\begin{aligned} F_p &\simeq (\pi/2)\ \ (64/11), \\ &\simeq 9{\cdot}14. \end{aligned}$$

Substituting for F_p in Eq. 1 gives:

$$\begin{aligned} V_{CERM} &\simeq 0{\cdot}85 \,.\, 9{\cdot}14 \,.\, 150\text{V}, \\ &\simeq 1170\text{V}. \end{aligned}$$

The calculation neglects the voltage drop across the flash-over protection resistor R_{28}. This reduces the average d.c. voltage to the timebase by 14·5V; thus the true $V_{h.t.}$ supply is 135·5V, this renders a corrected V_{CERM} value of 1050V.

The deflection coil VA is calculated from:

$$VA = (0{\cdot}85 \,.\, F_p \,.\, L_y \,.\, I_y^2)/\tau_s, \qquad \ldots(3)$$

where L_y (deflection coil inductance) = 2·1mH at 1kHz,
I_y (deflection coil current) = 3A pk-pk (see Fig. 6c),
$\tau_s = \tau_t - \tau_f$ = 53μs.

Substituting in Eq. 3 gives:

$$\begin{aligned} \text{deflection coil VA} &= (0{\cdot}85 \,.\, 9{\cdot}14 \,.\, 2{\cdot}1 \,.\, 10^{-3} \,.\, 3^2)/53 \,.\, 10^{-6}, \\ &= 2770\text{VA} \end{aligned}$$

The transistor TR_4 must be capable of handling the deflection coil VA plus an additional 15% for linearity correction and 5% for shunting by the line output transformer (T_1), and minus 5% for S correction; that is: 2770 . 1·15 = 3180VA.

The recovery efficiency (the ratio of the reverse collector current to the forward collector current) is approximately 70% (see Fig. 6a). Thus TR_4 will switch 3180 × 1/(1 + 0·7), or 1870VA, in the forward direction.

$$\begin{aligned} I_{CM} \text{ nominal} &= VA/V_{CERM} \\ &= 1870/1050 \\ &= 1{\cdot}78\text{A}. \end{aligned}$$

The V_{CERM} and I_{CM} values calculated are nominal. By making allowances of 20% for operational spreads and component tolerances, and 10% for unstabilised operation of the line output stage, an increase of 30% can be expected. Thus V_{CERM} is increased to 1365V and I_{CM} to 2·32A, and TR_4 is then required to handle 3160VA. This is within the VA rating of the BU105 transistor. Hence by selecting the nominal design value calculated for V_{CERM} and I_{CM}, the BU105 operates safely under all contingencies.

Figs. 6a and 6b show the BU105 I_C and V_{CE} waveforms, and Figs. 6c and 6d the deflection coil I_y and V_y waveforms. Note that the voltage waveforms show that the line output transformer T_1 (development type DT2048) is third harmonic tuned.

Low voltage supplies

Two l.t. supplies of 26V d.c. and 12V d.c. are derived from the scan rectification winding on T_1 and provide 20W of l.t. power for the field scan, audio output and other circuits in

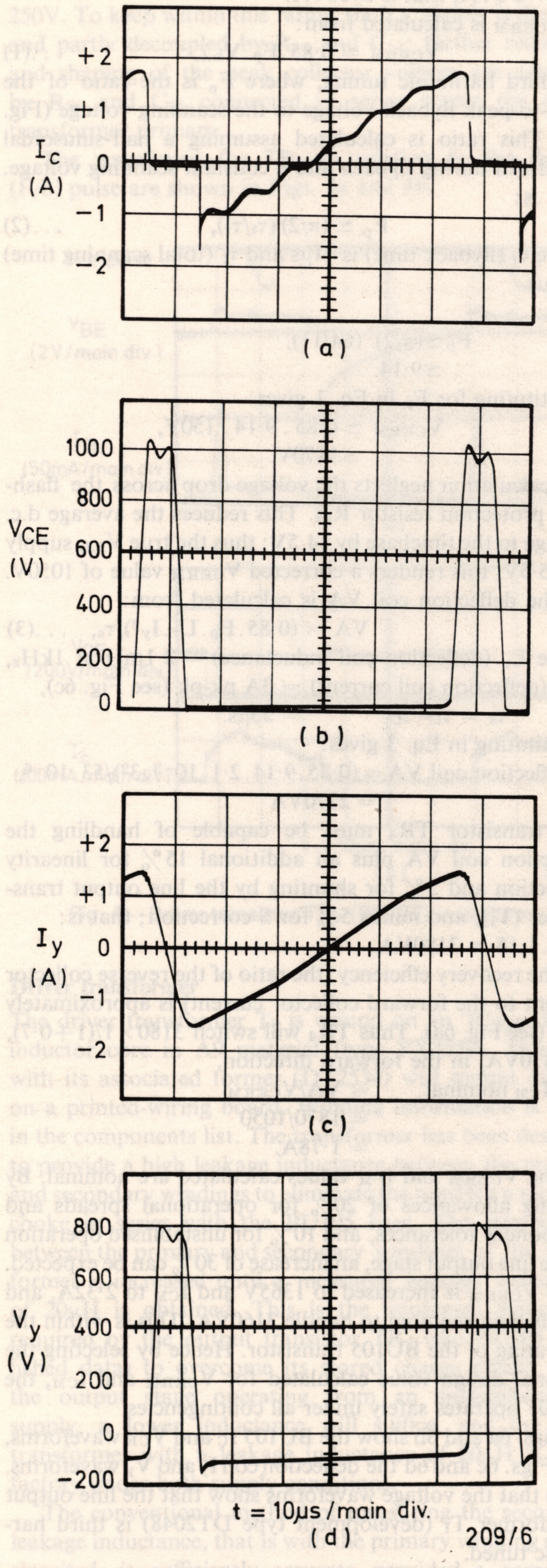

Fig. 6—Line output stage waveforms:
(a) transistor TR_4 (BU105) collector current
(b) transistor TR_4 (BU105) collector voltage
(c) deflection coil current
(d) deflection coil voltage

the receiver. Since sufficient power is available from this winding, the shunt stabiliser circuit, described in Ref. 1, can be used.

During the scanning period the rectifier diode D_5, a BYX55, and the low resistance BU105 transistor are conducting. It follows that the source impedance for the l.t. supplies depends largely on the source impedance of the 150V supply. This is about 56Ω given by the protection resistance R_{28}; when reflected across the scan rectification winding it provides a low impedance source for the l.t. supplies. Scan rectification will consequently make no difference to the flyback peak voltage or result in any appreciable change to the peak collector current of TR_4.

A_1 and focus supplies

A tapping point on T_1 provides a voltage source for the A_1 and focus supplies for the tube. Rectification of the flyback pulse is achieved by a BY184 silicon diode D_4.

Picture tube e.h.t.

The picture tube e.h.t. of 18kV is developed by a BY185 silicon diode D_3. The e.h.t. regulation characteristic is shown in Fig. 7.

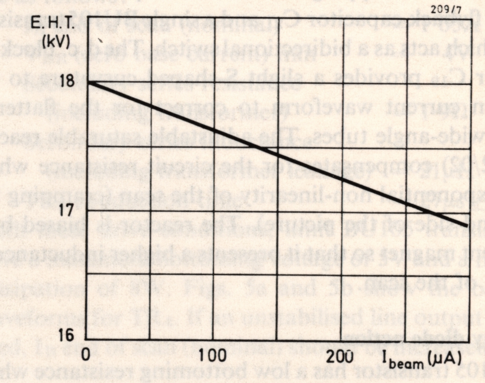

Fig. 7—Picture tube e.h.t. regulation

Measurements

Measurements on the line output stage operating with the shunt stabiliser circuit (Ref. 1) are given in Table 2.

FLASHOVER PROTECTION

During picture tube flashover, high voltage pulses are generated (Refs. 3 and 4). These appear in the circuit via the capacitance between the deflection coils and the inside conductive layer of the picture tube. The line output transistor, TR_4, is protected against these pulses by an undecoupled 56Ω resistor (R_{28}) connected in series with the h.t. supply. Further protection against flashover is obtained by taking the emitter of TR_4 and one side of the deflection coils to a common earth.

To ensure correct operation of the line timebase, the tuning capacitor C_{21} must be connected directly across the collector and the emitter of TR_4. Similarly, both the earthy end of the driver transformer secondary winding, and one end of the CR circuit in the base, should be connected directly to the emitter. Thus the latter should serve as the common earthing point.

The 1nF capacitor C_{20} gives the base-emitter diode of TR_4 protection against transient voltages which may appear on the collector and be transmitted to the base through the reverse-biased base capacitance. Capacitor C_{20} should be mounted close to the base and emitter terminals of TR_4.

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TABLE 2
Measurements on the line output stage operating with the shunt stabiliser circuit

		I_{beam} 0µA	I_{beam} 300µA	
	$V_{h.t.}$	150	148	V
	$I_{h.t.}$	260	270	mA
	I_y pk–pk	3·0	3·0	A
	t_f (flyback time)	11·2	11·3	µs
	e.h.t.	18	16·9	kV
BU105	V_{CERM}	1050	1050	V
(TR_4)	$+I_{CM}$	1·70	1·65	A
	$-I_{CM}$	1·15	1·10	A

Low voltage supplies: $V_{lt} = 26{\cdot}5$v, $I_{lt} = 740$mA.

COMPONENTS LIST

Capacitors

		Mullard type
C_1, C_{13}	80µF 25V	C426 AR/F80
C_2	47nF	C280 AE/P47K
C_3, C_{15}	3·3nF	C296 AC/A3K3
C_4, C_{11}	10µF 64V	C426 AR/H10
C_5	220nF	C280 AE/P220K
C_6, C_9, C_{12}, C_{14}	100nF	C280 AE/P100K
C_7	1µF	C280 AE/A1M
C_8	2·2nF 5%	424 42202
C_{10}	8·2nF 5%	C295 AH/D8K2
C_{16}	1·5nF	C296 AC/A1K5
C_{17}	100nF 400V	C280 CF/P100K
C_{18}	220pF	C333 CH/C220E
C_{19}	270nF 250V	
C_{20}	1nF	C296 AC/A1K
C_{21}	2·4nF 5% 1·5kVmax	278 72242
C_{22}	1µF 500V	
C_{23}	100nF 500V	
C_{24}	10nF 500V	
C_{25}	16µF 30V	

Diodes		Transistors	
D_1, D_2	BA154	TR_1, TR_2	BC158
D_3	BY185	TR_3	BF337
D_4	BY184	TR_4	BU105
D_5	BYX55		

BU105 heatsink

80cm² of blackened aluminium of thickness 3.0mm.

Resistors

All resistors are in the CR16 range except:

*R_{18}	12kΩ	5W
*R_{19}	3·9kΩ	5·8W max
*R_{20}	4·7kΩ	1·5W max
*R_{23}	1·5Ω 5%	1·5W
R_{28}	56Ω	5·0W

---

*For an unstabilised 150V supply to the line output stage, R_{20} is 3·3kΩ, 1·5W, and R_{23} 1Ω, 1W. For a thyristor 150V stabilised supply, R_{18} and R_{19} are 5·6kΩ, 8W, and 820Ω, 2W respectively.

WOUND COMPONENTS

Deflection coils

Mullard AT1040	Inductance at 1kHz:	2·1mH
	Resistance at 25°C:	3·9Ω
	Pk-pk current for edge to edge scan at 18kV:	2·82A

Linearity inductor

Mullard AT4042/02

Oscillator coil (linehold)

The coil is wound on a ¼in (6·35mm) Aladdin short former.

Winding—2000 turns, wave wound 10mm wide with 0·8mm (44 swg) cu. polyurethane wire grade 1. Tapping point T is at 500 turns from S.

Gears on Douglas wave winder

A	B	C	D	E	F
40	35	44	50	24	96

Core—A ferrite rod or tube, 4mm diameter, 16mm long, in threaded plastic moulding.

Driver transformer

Core—FX3550

Former—DT2534

Primary winding—1000 turns of 0·14mm en. cu. wire.

Secondary winding—40 turns of 0·4mm en. cu. wire (1½ layers).

Primary-secondary interleaving—9 turns, 60µm, Makrofol N or Melinex.

Air gap—0·1mm spacers in centre and outside legs.

Secondary leakage inductance—bridge measurement with primary short circuited, 23µH.

Cold resistance: primary = 50Ω, secondary = 0·4Ω.

Line output transformer

This is an experimental design; development type DT2048.

Width control

Winding—100 turns of 0·315mm wire on a 4·5mm former. A Ferroxcube tube FX1054 is used for adjustment.

Transistor 90° colour line deflection and e.h.t. circuit

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Mullard Central Application Laboratory

UDC 621.397.334 [621.382.3]

The line timebase uses a single BU108 transistor rated at 1500V and 5A. The line output transformer overwind is tuned to the 5th harmonic and a voltage trebler is used to obtain 25kV for the picture tube. The transistor oscillator, driver and flywheel stages are described and 20 and 40V supplies can be obtained by scan rectification of the line output transformer waveform.

INTRODUCTION

In order to scan a 90° colour tube at 25kV the absolute maximum peak volt-ampere rating of the output transistor must be about 4500VA. A BU108 transistor, rated at 1500V and 5A is capable of 7500VA, and therefore a one transistor line timebase has been designed around the BU108. This transistor can be operated directly from an h.t. line of 148V, and low voltage supplies for other circuits can be obtained by scan rectification of the line output transformer waveform. The circuit is shown in Fig. 1. A thyristor power supply is recommended, the design principles of which are discussed in Ref. 1.

LINE OUTPUT STAGE

In order to allow for component spreads and operational variations it is necessary to allow an increase above the nominal of 20% on the peak values, I_{CM} and V_{CERM}, of the transistor. Since the absolute maximum V_{CERM} of the BU108 is 1500V, the nominal value must be ≤1250V. With a 11·2μs flyback time, t_f, and 64μs total time t_t,

$$V_{CERM} = \pi t_t V_{ht}/2t_f,$$

where V_{ht} is the supply voltage, and V_{CERM} is the peak collector-emitter voltage, and $R_{BE} = 100\Omega$.

Therefore $V_{ht} = V_{CERM}\times 2t_f/\pi t_t$

$$= 1250\times 2\times 11{\cdot}2/64\pi = 139V.$$

The actual h.t. supply is 148V because of the voltage dropped across R_{26}, a 27Ω flashover protection resistor, which is in series with the h.t. rail. The value of the h.t. supply would have to be reduced if no l.t. supplies were obtained from the timebase since the voltage drop across R_{26} would be less.

The peak current and peak voltage that the transistor has to handle is made up of the peak VA of the scanning coils, plus the VA for scanning correction, for the convergence circuit, and for the shunting by the line transformer through its finite inductance.

This additional VA is approximately 40% of the scanning coil VA, made up of 20% for linearity correction, 10% for transformer shunting and 7% for convergence. The switched VA is then 1·4 times the scanning coil VA.

The scanning coil VA $= F_pL_y(I_y)^2/(t_t - t_f)$

where $F_p = \pi t_t/2t_f$, L_y is the scanning coil inductance, and I_y the current flowing through the scanning coils.

With an 11·2μs flyback time, an L_y of 2·8mH, and an I_y of 2·8A,

$$\text{switched VA} = \frac{1{\cdot}4\times\pi\times 64\times10^{-6}\times 2{\cdot}8\times10^{-3}\times(2{\cdot}8)^2}{2\times 11{\cdot}2\times10^{-6}\times 52{\cdot}8\times10^{-6}}$$

$$= 5220\text{VA}.$$

For a typical timebase the peak current at the start of a scan is 70% of the value at the end of a scan (recovery efficiency). Therefore:

transistor VA = 5220/1·7 = 3070VA.

For an h.t. line of 139V where V_{CERM} = 1250V, I_{cm} = 3100/1250 = 2·46A.

The calculated I_{cm} is a nominal value, and if it is increased by 20% to allow for component spreads, the absolute maximum value becomes I_{cm} = 3·0A. As the BU108 has an absolute maximum current rating of 5A, there are ample safety margins.

LINE OUTPUT TRANSFORMER AND E.H.T. GENERATION

The line output transformer windings are as shown in the circuit diagram (Fig. 1). Details of the windings are given in the components list. The overwind is wound as a continuation of the primary, but separated from it by two layers of interleaving. A small variable inductor L_3 is connected in series with the overwind, to obtain 5th harmonic tuning. The tuning was optimum when the 8·4kV pulse at the input to the trebler is as flat as possible (Figs. 2 and 3). Fifth harmonic tuning gives a flat-topped input pulse to a trebler and therefore improved e.h.t. regulation. The regulation is further improved by means of a diode clamping

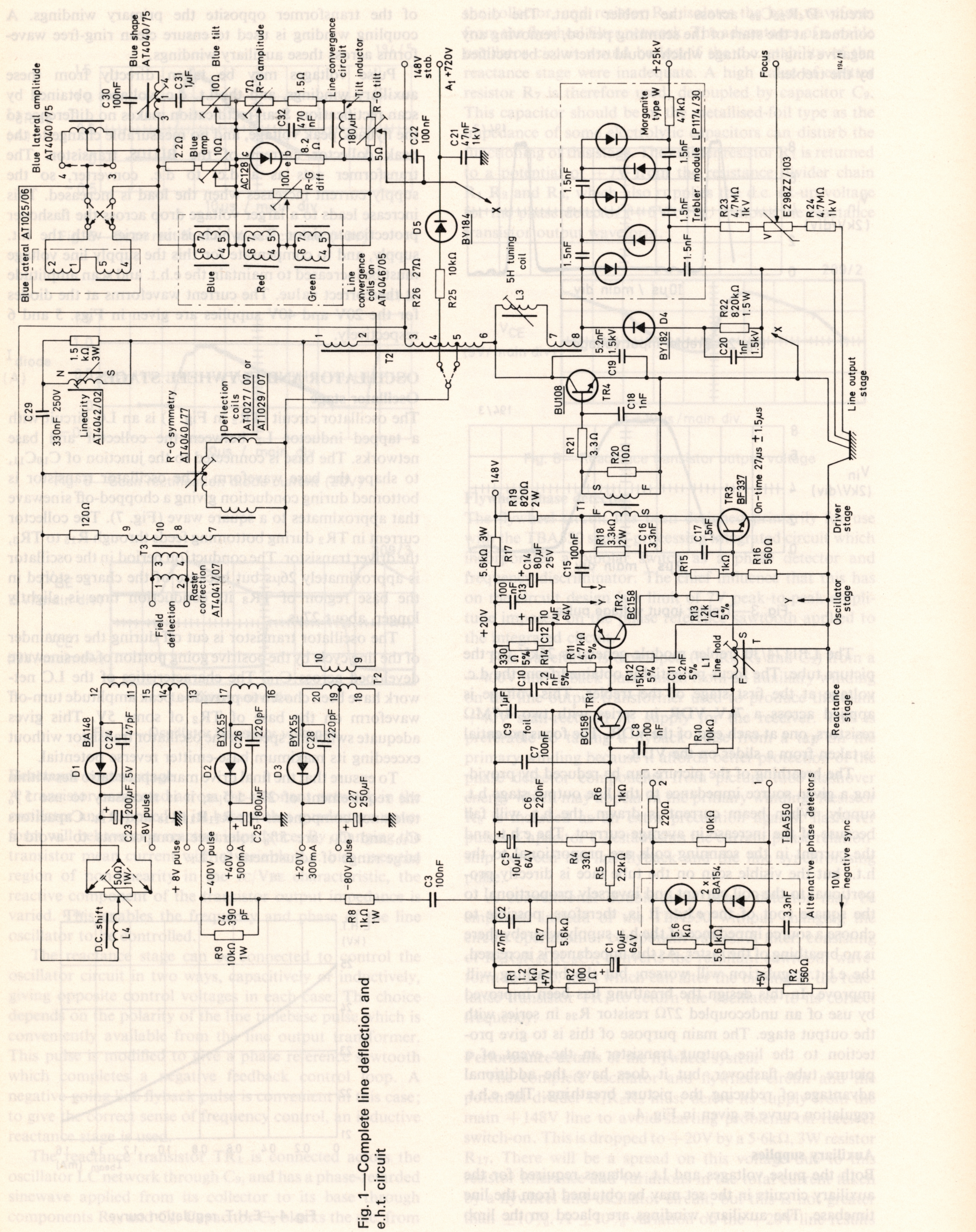
Blue lateral AT1025/06
Blue lateral amplitude AT4040/75
Blue shape AT4040/75
Blue amp
Blue tilt
R-G amplitude
R-G diff
Line convergence circuit
Tilt inductor 180μH
R-G tilt
Line convergence coils on AT4046/05
Blue
Red
Green
148V
A1 +720V
Focus
+25kV
Morganite type W
Trebler module LP1174/30
E298ZZ/103
BY184
BY182
D4
D5
5H tuning coil
L3
T1
T2
BU108
TR4
Line output stage
Driver stage
TR3 BF337
On-time 27μs ±1.5μs
Oscillator stage
L1 Line hold
Reactance stage
BC158
TR1
TR2
TBA550
Alternative phase detectors
10V negative sync
+20V
+5V
Field deflection
Raster correction AT4041/07
Linearity AT4042/02
R-G symmetry AT4040/77
Deflection coils AT1027/07 or AT1029/07
D.C. shift
BA148
BYX55
D1
D2
D3
+8V pulse
−8V pulse
−400V pulse
+40V 500mA
+20V 300mA
−80V pulse

Fig. 1—Complete line deflection and e.h.t. circuit

circuit $D_4R_{22}C_{20}$ across the trebler input. The diode conducts at the start of the scanning period, removing any negative ringing voltage which would otherwise be rectified by the trebler.

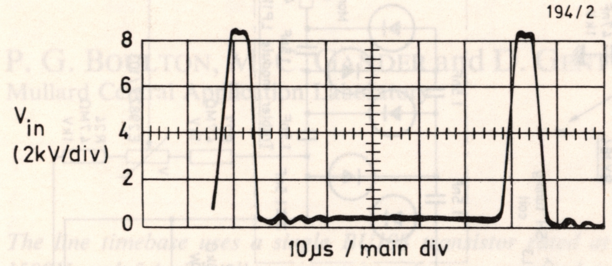

Fig. 2—Trebler input voltage

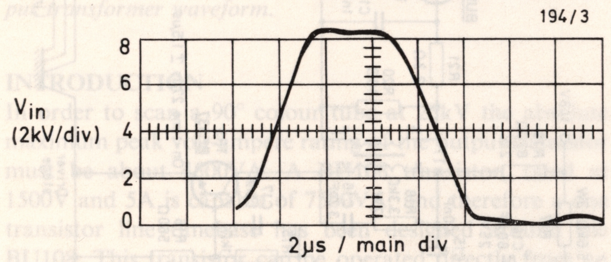

Fig. 3—Trebler input voltage pulse

The LP1174/30 trebler module generates 25kV for the picture tube. The focus potential is obtained from the d.c. voltage at the first stage of the trebler. This voltage is applied across a 7kV VDR in series with two 4·7MΩ resistors, one at each end of the VDR. The focus potential is taken from a slider on the VDR.

The breathing of the picture can be reduced by providing a given source impedance to the line output stage h.t. supply. When beam current is drawn, the h.t. will fall because of the increase in average current. The e.h.t. and the current in the scanning coils are proportional to the h.t., but the visible scan on the tube face is directly proportional to the coil current and inversely proportional to the square root of the e.h.t. It is, therefore, possible to choose a source impedance of the h.t. supply whereby there is no breathing of the raster. As this impedance is increased, the e.h.t. regulation will worsen, but the breathing will improve. In this design the breathing has been improved by use of an undecoupled 27Ω resistor R_{26} in series with the output stage. The main purpose of this is to give protection to the line output transistor in the event of a picture tube flashover, but it does have the additional advantage of reducing the picture breathing. The e.h.t. regulation curve is given in Fig. 4.

Auxiliary supplies

Both the pulse voltages and l.t. voltages required for the auxiliary circuits in the set may be obtained from the line timebase. The auxiliary windings are placed on the limb of the transformer opposite the primary windings. A coupling winding is used to ensure clean ring-free waveforms across these auxiliary windings.

Pulse voltages may be taken directly from these auxiliary windings, and the l.t. d.c. voltages obtained by scan rectification. Scan rectification makes no difference to the flyback peak voltage, and no measurable change to the peak collector current of the BU108 transistor. The transformer acts as a d.c. to d.c. converter, so the supply current increases when the load is increased. This increase leads to a larger voltage drop across the flashover protection resistor R_{26} which is in series with the h.t. supply, and to compensate for this the supply line voltage must be increased to maintain the e.h.t. and scan amplitude at the correct value. The current waveforms at the diodes for the 20V and 40V supplies are given in Figs. 5 and 6 respectively.

OSCILLATOR AND FLYWHEEL STAGES

Oscillator stage

The oscillator circuit (TR_2 in Fig. 1) is an LC circuit with a tapped inductor L_1 between the collector and base networks. The base is connected to the junction of $C_{10}C_{11}$, to shape the base waveform. The oscillator transistor is bottomed during conduction giving a chopped-off sinewave that approximates to a square wave (Fig. 7). The collector current in TR_2 during bottoming feeds through R_{15} to TR_3, the driver transistor. The conduction period in the oscillator is approximately 26μs but because of the charge stored in the base region of TR_3 its conduction time is slightly longer, about 27μs.

The oscillator transistor is cut off during the remainder of the line cycle by the positive going portion of the sinewave developed across C_{10}. The characteristics of the LC network have been chosen to provide a peak amplitude turn-off waveform on the base of TR_2 of some 5V. This gives adequate switching speed in the oscillator transistor without exceeding its maximum base-emitter reverse potential.

To ensure that the final drive mark-space ratio lies within the requirement of 27±1·5μs, it is necessary to use 5% tolerance components for $R_{11}R_{12}R_{13}$ and R_{14}. Capacitors C_{10} and C_{11} are 5% tolerance components to avoid a large range of adjustment for L_1.

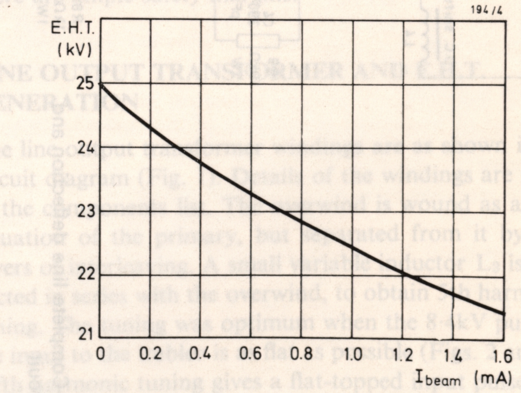

Fig. 4—E.H.T. regulation curve

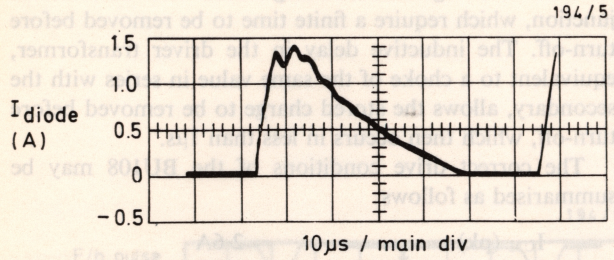

Fig. 5—Scan rectification diode current (40V)

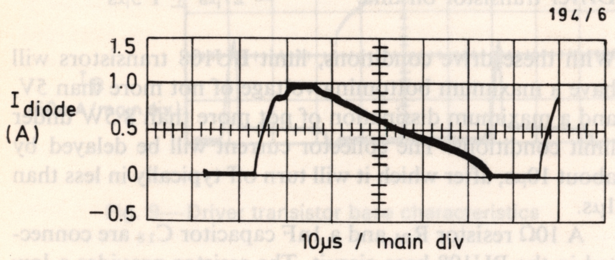

Fig. 6—Scan rectification diode current (20V)

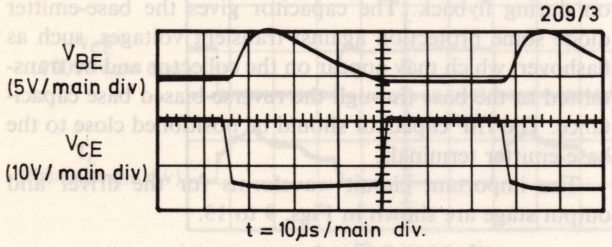

Fig. 7—Oscillator transistor voltages

Reactance control stage

A transistor connected across the resonant network in the oscillator and supplied at its base with a quadrature waveform will behave as a reactive impedance. By changing the transistor mean current with a variable d.c. bias over the region of non-linearity in the I_C/V_{BE} characteristic, the reactive component of the transistor output impedance is varied. This enables the frequency and phase of the line oscillator to be controlled.

The reactance stage can be connected to control the oscillator circuit in two ways, capacitively or inductively, giving opposite control voltages in each case. The choice depends on the polarity of the line timebase pulse which is conveniently available from the line output transformer. This pulse is modified to give a phase reference sawtooth which completes a negative feedback control loop. A negative-going line flyback pulse is convenient in this case; to give the correct sense of frequency control, an inductive reactance stage is used.

The reactance transistor TR_1 is connected across the oscillator LC network through C_9, and has a phase-retarded sinewave applied from its collector to its base through components R_{10} and C_8. Capacitor C_8 blocks the d.c. from the collector, and resistor R_{10} isolates the base waveform from the flywheel filter network. The advantages of a stable oscillator circuit would be lost if the d.c. stability of the reactance stage were inadequate. A high value of emitter resistor R_7 is therefore used, decoupled by capacitor C_9. This capacitor should be of the metallised-foil type as the impedance of some electrolytic capacitors can disturb the functioning of this stage. The emitter resistor R_1 is returned to a potential of +7V from the resistance divider chain R_1 R_2 and R_3, which also supplies the d.c. sit-up voltage for the phase detector (+5V). Fig. 8 shows the reactance transistor output waveform.

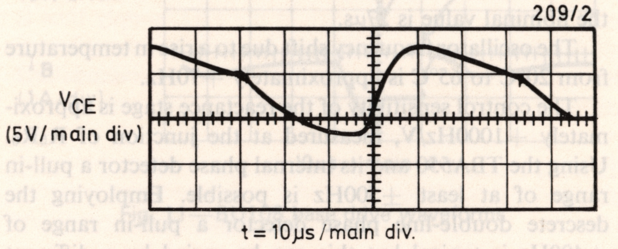

Fig. 8—Reactance transistor output voltage

Flywheel phase detector

The flywheel circuit has been designed primarily for use with the TBA550 signal-processing integrated circuit which incorporates a bipolar switch as a phase detector and frequency discriminator. The chief influence that this has on the circuit design is a limit of 7V peak-to-peak amplitude imposed on the phase reference sawtooth applied to the integrated circuit.

This waveform is developed (by R_8 R_9 and C_2) from a −400V line flyback pulse taken from an auxiliary winding on the line output transformer used to produce the main +40V scan-rectified l.t. supply for the receiver. This is preferable to taking a flyback pulse from a tap on the primary winding because it affords better protection of the phase detector components from picture tube flashover energy which may appear in the primary winding. Resistor R_8 is included as an extra precaution against flashover pulses. Capacitor C_4 establishes the current phase relationship between the line flyback and the line synchronising pulse.

A conventional double-diode phase detector may be employed instead of the TBA550 without changing the circuit operation or component values. A filter, consisting of $R_4R_5R_6C_5C_6C_7$, converts the resultant sawtooth waveform to a d.c. level which can alter the biasing of the reactance transistor TR_1 to return the oscillator to its correct frequency.

Performance details of the flywheel system

The complete oscillator and flywheel circuit and the potential divider $R_1R_2R_3$ must derive its supply from the main +148V line to avoid starting problems on receiver switch-on. This is dropped to +20V by a 5·6kΩ, 3W resistor R_{17}. There will be a spread on this voltage due to this resistor tolerance and variations in the total current taken by a flywheel and oscillator circuit, but this is not greater than ±10%. A ±10% variation on the +20V line results

in an oscillator frequency shift of some ±170Hz. However, because the +148V line is stabilised no compensation for this effect is necessary.
The total spread in conduction time of the line driver transistor is the sum of the rms values of the tolerances of the critical components.

A summary of the results is as follows:

Statistical sum of the effects of all the passive component tolerances:	±0·93μs
TR_2 tolerance	±0·55μs
TR_3 tolerance	±1·0μs
total rms value	±1·47μs

the nominal value is 27μs.

The oscillator frequency shift due to a rise in temperature from 20°C to 65°C is approximately +70Hz.

The control sensitivity of the reactance stage is approximately +1000Hz/V, measured at the junction of R_5R_6. Using the TBA550 and its internal phase detector a pull-in range of at least ±500Hz is possible. Employing the descrete double-line phase detector a pull-in range of ±400Hz is typical but this can be varied by a different choice of flywheel filter components $R_4R_5C_5C_6$

To minimise the horizontal phase disturbance at the top of the picture due to the field pulse group a value of 33Ω for R_4 is necessary for the TBA550 but for the double diode circuit 47Ω is more suitable.

An aspect of practical design which should be mentioned is the importance of avoiding any time-constants in the flywheel and oscillator system which would cause an excessively long "climb-up" from a possibly low oscillator frequency to the correct frequency immediately following receiver switch-on. Such an occurrence could endanger the line output transistor because of the high peak voltage and currents. The time constants of the flywheel system, mainly that of the electrolytic capacitor, C_5, have been arranged so that a minimum delay occurs before the flywheel system becomes synchronised.

The coil design of L_1 is intended for use without a screening can. If in a particular receiver layout a can is necessary the number of turns will have to be increased to compensate for the drop in inductance.

DRIVER STAGE

The driver stage is designed to operate from the 148V stabilised supply. The BF337, with a peak collector voltage rating of 250V, is suitable for this application. The h.t. is dropped and partially decoupled by an 820Ω resistor, R_{19}, and a 100nF capacitor, C_{15}. Further reduction and shaping of the peak collector voltage is obtained by a 4·7kΩ resistor R_{18} and a 3·3nF capacitor C_{16} across the driver transformer primary winding. The peak and average collector currents of the BF337 are 110mA and 36mA when the collector conduction time is 27μs. The driver transformer is wound on a FX3550 ferrite core. The secondary is separated from the primary by nine turns of interleaving which gives a high leakage inductance and slows down the base current turn-off waveform.

This inductance accommodates stored charge effects as, to ensure that a limit BU108 is bottomed during scan, a base current is supplied which is higher than that required for nominal transistors. This base overdrive leads to large numbers of charge carriers being stored near the collector junction, which require a finite time to be removed before turn-off. The inductive delay in the driver transformer, equivalent to a choke of the same value in series with the secondary, allows the stored charge to be removed before turn-off, which then occurs in less than 1μs.

The correct drive conditions of the BU108 may be summarised as follows:

I_{CM} (pk) = 2·6A
I_B (end-of-scan) = 0·7A
V_{BE} (zero base current) = −3V minimum
Transformer leakage inductance = 23μH
Driver transistor on time = 27μs ± 1·5μs

With these drive conditions, limit BU108 transistors will have a maximum bottoming voltage of not more than 5V, and a maximum dissipation of not more than 8·5W under limit conditions. The collector current will be delayed by about 10μs, after which it will turn off typically in less than 1μs.

A 10Ω resistor R_{21} and a 1nF capacitor C_{18} are connected in the BU108 base circuit. The resistor provides a low impedance path when the BU108 is turned off. Without this resistor, the base circuit is lightly loaded, and ringing oscillations are likely to occur, which may turn the BU108 on during flyback. The capacitor gives the base-emitter diode some protection against transient voltages, such as flashover, which may appear on the collector and be transmitted to the base through the reverse-biased base capacitance. The 1nF capacitor should be positioned close to the base-emitter terminals.

The important circuit waveforms for the driver and output stage are shown in Figs. 9 to 13.

TABLE 1
Measured values of main circuit parameters

I_{BEAM}	0	1·0	1·5	mA
V_{ht}	148	148	148	V
I_{ht}	430	580	640	mA
BU108				
V_{CERM}	1250	1190	1150	V
$+I_{CM}$(pk)	2·5	2·55	2·6	A
$-I_{CM}$(pk)	1·6	1·5	1·3	A
I_y(pk-pk)	2·8	2·7	2·6	A
*t_f	11·2	11·5	11·7	μs
picture breathing	0	1·7	2·0	%
e.h.t.	25·0	22·5	21·4	kV
Low voltage supplies				
V_{lt1}	40	38·7	37·8	V
I_{lt1}	510	494	482	mA
V_{lt2}	20	19·5	19·1	V
I_{lt2}	260	254	248	mA

If no l.t. supplies are used, the h.t. voltage must be reduced to 142 volts to obtain the correct e.h.t. and scan.
*This was measured at the bottom of the flyback waveform as shown on the oscilloscope.

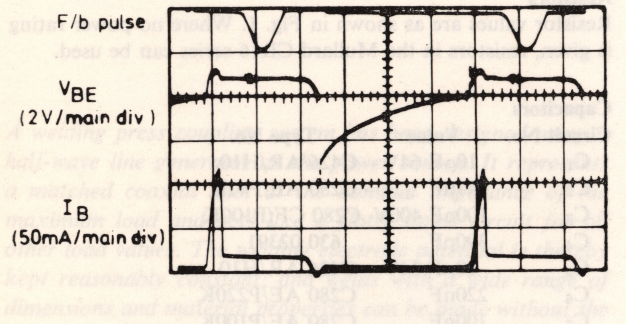

Fig. 9—Driver transistor base characteristics

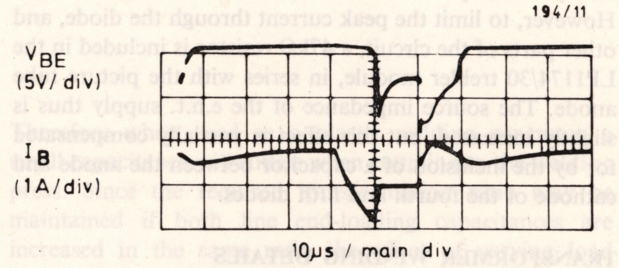

Fig. 11—BU108 base drive waveforms

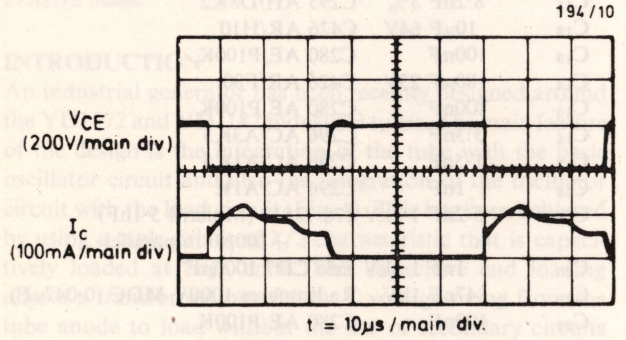

Fig. 10—Driver transistor output characteristics

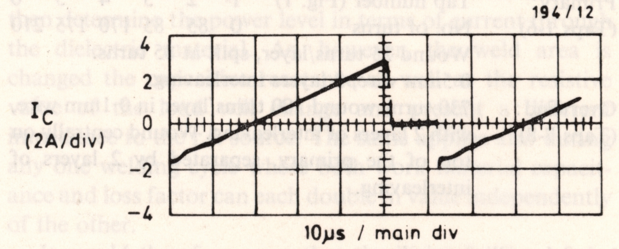

Fig. 12—BU108 scan rectification current

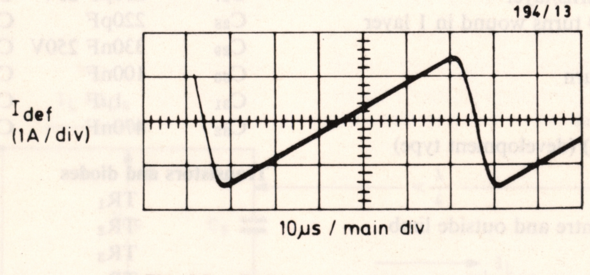

Fig. 13—Reflection coil current

FLASHOVER

During a flashover, a large transient is passed through the capacitance of the deflection coils to the inside conductive layer of the picture tube. Resulting voltages appearing on the line deflection coils can be harmful to semiconductor devices. For this reason, no pulses or timing signals should be taken from the winding on the line output transformer connected to the deflection coils.

The clamping diode circuit across the trebler input also provides protection for the BU108 during flashover. However, to limit the peak current through the diode, and other parts of the circuit, a 47kΩ resistor is included in the LP1174/30 trebler module, in series with the picture tube anode. The source impedance of the e.h.t. supply thus is slightly increased, but this effect is partially compensated for by the inclusion of a capacitor between the anode and cathode of the fourth and fifth diodes.

TRANSFORMER WINDING DETAILS

Line output transformer

Core: U core FX3187 I bar FX3188
Airgap: 2×0·3mm
Former: I.D. = 19·1mm, Thickness = 1·0mm
Length = 35mm
Wire: en. cu. PUR grade 1
Interleaving: 60μm Makrofol N

Primary:	Tap number (Fig. 1)	1	2	3	4	5	6
(Taps 1-6)	No. of turns	0	85	85	170	175	210

Wound 55 turns/layer, split at 85 turns.
0·5mm wire, 2 layers interleaving

Overwind: (Taps 7-8) 730 turns wound 100 turns/layer in 0·1mm wire, with 2 layers of interleaving. Wound centrally on top of the primary, separated by 2 layers of interleaving.

l.t. and auxiliary windings:

All windings in 0·4mm wire with 1 layer of interleaving
The windings should be wound in the order given

Taps 16-17 45 turns wound in 1 layer
„ 9-10 92 turns wound 46 turns/layer
„ 18-19-20 23 turns tapped at 10 turns wound in 1 layer
„ 11-12 2 turns
„ 13-14-15 2 turns tapped at 1 turn

Driver transformer

Core: FX3550 (A9 material) (development type)
Former: DT2534
Clips: DT2406
Airgap: 0·1mm spacers on centre and outside limb

Primary: 1000 turns of 0·14mm en. cu. wire
Secondary: 40 turns of 0·4mm en. cu. wire wound 26 turns/layer
Primary-Secondary Interleaving: 9 turns 60μm Makrofol N or Melinex giving a leakage inductance of 23μH

Oscillator coil (L_1)

Former: ¼ in (0·64mm) Aladdin short former
Winding: 2000 turns tapped at 500 turns of 0·8mm (44 swg) cu. PUR, grade 1 wire, wave wound 10mm wide. Gears on Douglas wave winder

A	B	C	D	E	F
40	35	44	50	24	96

Core: 4mm diameter ferrite rod or tube, 16mm long, in threaded plastic moulding

5H tuning coil (L_3)

Core: FX1054 in threaded plastic moulding
Former: ¼ in (0·64 mm) Aladdin short former
Winding: 400 turns wound 100 turns/layer
Wire: 0·1mm en. cu.
Interleaving: 1 layer of 60μm Makrofol N

COMPONENTS

Resistors

Resistor values are as shown in Fig. 1. Where no power rating is given, resistors in the Mullard CR16 series can be used.

Capacitors

Circuit No.	Value	Type No.
C_1	10μF 64V	C426 AR/H10
C_2	47nF	C280 AE/P47K
C_3	100nF 400V	C280 CF/P100K
C_4	390pF	630 02391
C_5	10μF 64V	C426 AR/H10
C_6	220nF	C280 AE/P220K
C_7	100nF	C280 AE/P100K
C_8	100nF	C280 AE/P100K
C_9	1μF	C280 AE/A1M
C_{10}	2·2nF 5%	424 42202
C_{11}	8·2nF 5%	C295 AH/D8K2
C_{12}	10μF 64V	C426 AR/H10
C_{13}	100nF	C280 AE/P100K
C_{14}	80μF 25V	C426 AR/F80
C_{15}	100nF	C280 AE/P100K
C_{16}	3·3nF	C296 AC/A3K3
C_{17}	1·5nF	C296 AC/A1K5
C_{18}	1nF	C296 AC/A1K
C_{19}	5·2nF 1·5kV	278 72512 (Mullard 5·1nF) +100pF in parallel
C_{20}	1nF 1·5kV	Erie CD7 1000pF
C_{21}	47nF 1kV	Radiospares 1000V MDC (0·047μF)
C_{22}	100nF	C280 AE/P100K
C_{23}	200μF 6·4V	C426 AR/C200
C_{24}	47pF	C333 CB/C47E
C_{25}	800μF 64V	C431 BR/H800
C_{26}	220pF	C333 CB/C200E
C_{27}	250μF 25V	C437 AR/F250
C_{28}	220pF	C333 CB/C200E
C_{29}	330nF 250V	C296 AA/A330K
C_{30}	100nF	C280 AE/P100K
C_{31}	1μF	C280 AE/A1M
C_{32}	470nF	C280 AE/A470K

Transistors and diodes

TR_1	BC158
TR_2	BC158
TR_3	BF337
TR_4	BU108
TR_5	AC128
D_1	BA148
D_2	BYX55
D_3	BYX55
D_4	BY182
D_5	BY184

REFERENCE

1. King, D. J., 'Thyristor power supplies for colour tv receivers'. **Mullard Technical Communications, 11,** 102 Nov. 1969, pp. 29 to 34.

Welding press coupling to a half-wave generator

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UDC 621.365.55.[678]: 678 [621.791]

A welding press coupling system has been designed using a half-wave line generator as the power source. It represents a matched coaxial duct at the nominal impedance of the maximum load and acts like a flatly tuned circuit for all other load values. The welding electrode potential is thereby kept reasonably constant, and welds with a wide range of dimensions and material properties can be made without the need for resetting of the r.f. power coupling system. There are good possibilities of operating such a welder within the limits set by the international interference regulations for the 27MHz band.

INTRODUCTION

An industrial generator has been recently designed around the YD1172 and YD1182 industrial tubes. The main feature of the design is the integration of the tube with the basic oscillator circuit and also the integration of the oscillator circuit with the load circuit (Fig. 1). This has been achieved by using a tank circuit of $\lambda/2$ characteristic that is capacitively loaded at both ends. This capacitive end loading allows a transformation of the r.f. voltage swing from the tube anode to load without the use of secondary circuits which might introduce coupling difficulties or promote parasitic oscillations at their own resonant frequency. The transformation facility that is given by the use of this system makes very high load position potentials possible. Therefore when used as a plastic welding generator a small capacitance will suffice when coupling to the welding press. Since the required transformation ratio may be maintained if both line end-loading capacitances are increased in the same ratio the effect of varying load coupling capacitances can be swamped to a large extent and improved frequency stability obtained.

CIRCUIT DESIGN

It is accepted that, irrespective of weld area an r.f. potential of around 1kV rms is needed across the electrodes to produce a satisfactory weld. The size of the weld area will then determine the power level in terms of current through the dielectric material. As, however, the weld area is changed the capacitive reactance as well as the resistive value of the load will change and present a different impedance to the r.f. source. The same applies also during any one welding cycle where both work material capacitance and loss factor can each double in value independently of the other.

It would therefore seem that the direct feeding of the welding electrodes from a coaxial system without stubs or other tuning arrangements is not possible since such a coaxial system would only deliver maximum power when terminated with a resistance of its own characteristic impedance value.

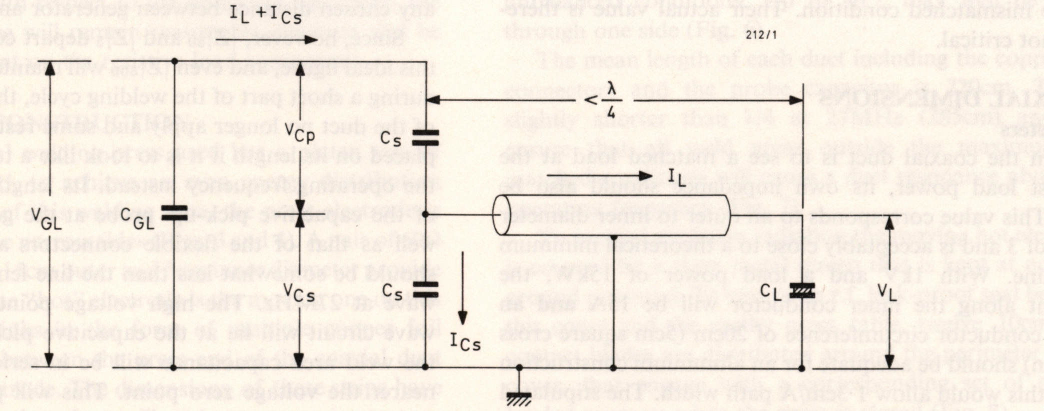

Fig. 1—Diagram of load coupling system between C_{GL}, the high potential end loading capacitance of the half-wave line generator and the press capacitance C_L. C_p is the pick-up probe capacitance and C_s represents the stray capacitance on the generator end of the duct.

There are, however, some possible modifications that will afford a workable compromise. The feeder line impedance can be so chosen that it will match reasonably well into the largest likely weld. The length of the feeder can be so dimensioned that, in the mismatched condition, it will look like a tuned circuit near the operating frequency that is heavily damped by the load.

Assessment of loading values

The half-wave generator built around the YD1172 tube delivers 15kW into a load. In plastic welding terms this means that, with two layers of 0·5mm p.v.c., it should be possible to fuse an area of 500cm^2 in a time of no more than six seconds. If necessary a more powerful generator could be made using the YD1182 tube.

For a smaller weld area of the same material both power level and timing are decreased to maintain a similar energy level per unit of weld area. When the reduction of weld area is at a ratio of 10:1 the absorbed power and welding time will be reduced by a factor of $1/\sqrt{10}$ to obtain a satisfactory weld. Values are given in Table 1.

TABLE 1
Loading values

Area (cm^2)	5	50	500
Welding time (s)	0·6	1·85	6
Power (kW)	1·5	4·65	15
Load resistance (Ω)	666·6	215	66·6
Capacitance (pF)	14·5	145	1447
Reactance (Ω)	406	40·6	4·06
Impedance (Ω) approx.	700	200	66

The capacitive value of the three areas can be calculated assuming two 0·5mm thicknesses of p.v.c. sheet of dielectric constant 6, and one 0·5mm thickness of Fibreglass reinforced plastic or Presspahn barrier material with a dielectric constant of 4.

$|Z|_{500}$ is very little different from the pure resistive value and $|Z|_{50}$ and $|Z|_{5}$ will be part of a flatly tuned circuit in the mismatched condition. Their actual value is therefore not critical.

COAXIAL DIMENSIONS

Diameters

If then the coaxial duct is to see a matched load at the highest load power, its own impedance should also be 66Ω. This value corresponds to an outer to inner diameter ratio of 3 and is acceptably close to a theoretical minimum loss line. With 1kV and a load power of 15kW, the current along the inner conductor will be 15A and an inner conductor circumference of 20cm (5cm square cross section) should be adequate for an aluminium construction since this would allow 1·3cm/A path width. The stipulated outer to inner diameter ratio of 3 will then give an outer diameter of 15cm. The use of square cross section duct for both inner and outer will not affect the calculated figures materially, but will ease the assembly (Fig. 2).

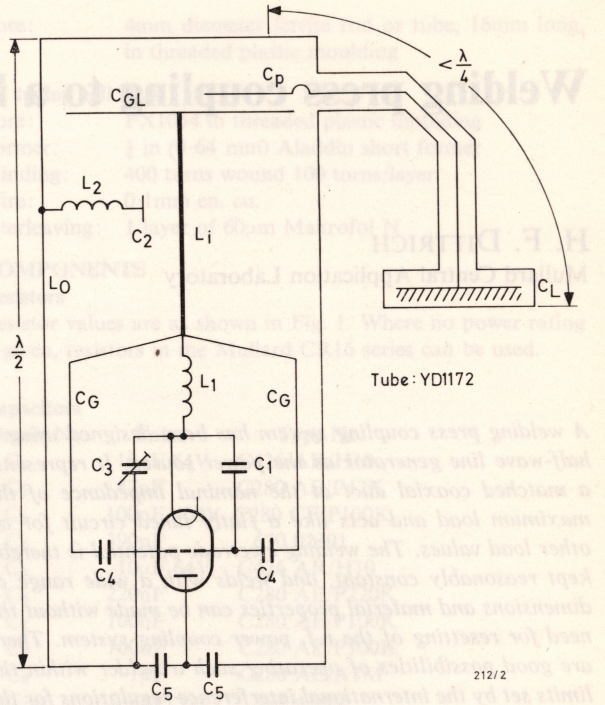

Fig. 2—Schematic diagram of the generator and single duct feeder to the plastic welding load. The components are as follows: C_G, generator tank circuit capacitance 200pF; C_{GL}, high potential end-loading capacitance 50pF; C_1, drive circuit loading capacitor 140pF; C_3, feedback capacitor 2 to 20pF; C_4, grid blocking capacitor 2·5nf (p.t.f.e. sandwich); C_5, anode blocking capacitor 1·5nF (p.t.f.e. sandwich); Li, 77Ω line section; L_0, line outer case; L_1, drive circuit inductor. The components L_2 and C_2 are for absorbing a secondary mode at 140MHz (Ref. 2)

Length of duct

Under matched conditions the length of the duct would, of course, be unimportant and could be made to traverse any chosen distance between generator and load.

Since, however, $|Z|_{50}$ and $|Z|_{5}$ depart considerably from this ideal figure, and even $|Z|_{500}$ will maintain its value only during a short part of the welding cycle, the line properties of the duct no longer apply and some restrictions must be placed on its length if it is to look like a tuned circuit near the operating frequency instead. Its length, including that of the capacitive pick-up probe at the generator end, as well as that of the flexible connectors at the press end, should be somewhat less than the line length of a quarter wave at 27MHz. The high voltage point of this quarter wave circuit will lie at the capacitive pick-up probe, and the weld area capacitance will be in series with this, but nearer the voltage zero point. This will prevent the duct frequency from coinciding with the generator frequency with very small weld areas. The generation of excessive potentials in the duct or at the welding electrodes is thereby avoided.

Probe size

The size of the capacitive pick-up probe can in the first place be determined by the full load requirements when the duct is assumed to be matched at the load end. In this condition the r.f. potential throughout the duct will be the same and equal to that required at the welding electrodes, namely the previously stated 1kV.

For the 15kW level the current in the duct will thus be 15A and since the source potential of the λ/2 generator is 12kV pk or 8·4kV rms the reactance of the probe capacitor will have to be:

$$X_{CP} = (8400–1000)/15 = 493\Omega$$

The probe capacitance is thereby:

$$C_p = 12\text{pF}.$$

Because of stray capacitance to ground provision should be made in the design to allow an increase of this value by about 20%.

Mismatched conditions

In the mismatched condition the ratio of the load current I_L over I_{CS}, the current through the stray capacitance, is so large that I_{CS} can be neglected for all practical purposes (Fig. 1). When, however, the load current I_L is decreased because of a change from $|Z|_{500}$ to $|Z|_5$, I_{CS} will become significant with respect to I_L and the capacitances dropping V_{CP} and V_{CS} will have to be seen as a potential divider from which the now tuned circuit duct is fed. The assumed probe capacitance is about 12pF, giving $X_{CP} = 500\Omega$, and the estimated stray capacitance is about 15pF, or $X_{CS} = 400\Omega$, paralleled by the load impedance of 780Ω. Thus a voltage division of 2:1 is obtained along this chain resulting in a probe potential of about 2·8kV rms. This is then also the potential on the high voltage point of the tuned circuit formed by the duct.

Any lower potential, such as the required 1kV at the welding electrodes, can then be tapped across a capacitor in series with the inductor (the line inner) of this circuit at the load position. Because the end loading capacitance and potential at the pick-up end of the circuit are relatively low and the circuit is heavily loaded as well as slightly off-resonance with respect to the operating frequency, the electrode voltage will remain reasonably constant and be largely dependent on the resistive load component.

PRACTICAL CONSTRUCTION

The professional welding press used has a platen size of 100×150cm and, to achieve an even energy distribution over the whole of this welding area, the press electrode is fed from both the narrow sides (Figs. 3 and 4). A pair of 60Ω coaxial ducts of 15cm outer and 5cm inner diameter provide the link. Since the "hot" electrode is the moving one on this press, flexible links in the form of multiple copper foil strips are used between the press end of the coaxial duct and the live electrode. The dimensions of these strips have been so chosen that they will at least approximate to a continuation of the duct impedance (Fig. 5).

At the generator end each coaxial terminates in a probe plate that has about 6pF capacitance with respect to the live plate of the generator line-end loading capacitor. Some flexibility has been provided in the mounting of these plates to allow for individual trimming. The r.f. potential distribution on the welding electrode can thereby be brought to an optimum. Alternatively these plates can also be linked to form a single 12pF capacitor to feed both ducts in parallel.

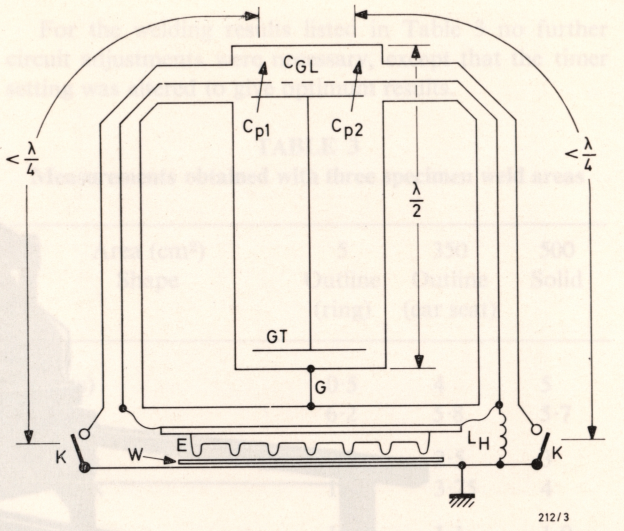

Fig. 3—Schematic diagram of twin duct press feed. The generator is the same as Fig. 2. C_{p1} and C_{p2} are the coupling probes, G are the grounding connections, K are the grounding contacts between the electrode cover and press table, E is the work electrode and W the press barrier material. L_H is a λ/4 choke, providing a d.c. ground potential for the coaxial inner. This is purely precautionary and will prevent the duct inner and electrode system assuming an unwanted charge

With two ducts in parallel the total matched load impedance would of course be shared; but since the platen itself has some inductance and welding patterns will rarely be symmetrically distributed, it is difficult to predict how the net load current will be shared under severely unbalanced conditions. Up to 90% may well be taken through one side (Fig. 6).

The mean length of each duct including the copper foil connectors and the probe capacitor is 220cm. This is slightly shorter than λ/4 at 27MHz (285cm) and will ensure that all weld areas outside the maximum for matched conditions will cause a duct resonance above the operating frequency (Fig. 7).

To prevent excessive radiation the moving hot electrode is covered by a sheet metal screen that is kept at nominal ground potential. To ensure an r.f. leak-proof seal between this cover and the earthy press table, twelve 10cm wide contact blades are distributed around the perimeter of the cover, that engage with a corresponding set of spring-loaded contacts when the press is closed (Fig. 8).

Additional grounding of the press table is obtained by a 50cm wide copper sheet that connects the press table to the outer skin of the λ/2 generator box.

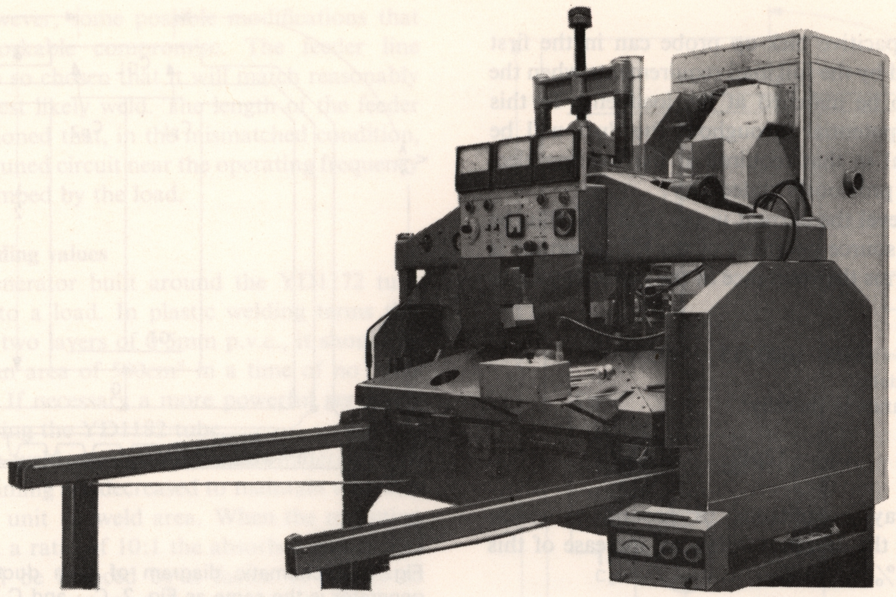

Fig. 4—Plastic welding press with feeder ducts and half-wave line generator at the rear.

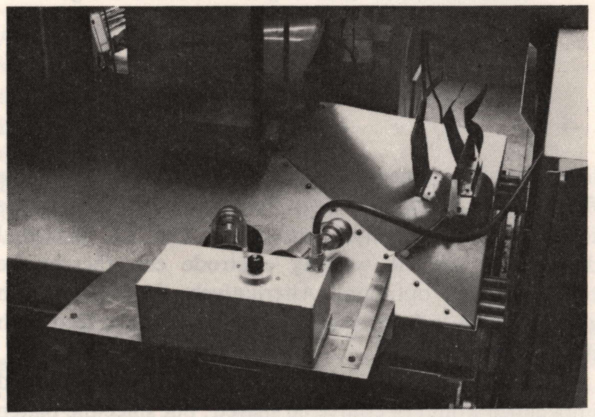

Fig. 5—Feeder ducts. Below them the copper sheet grounding connection between the press head and generator can be seen

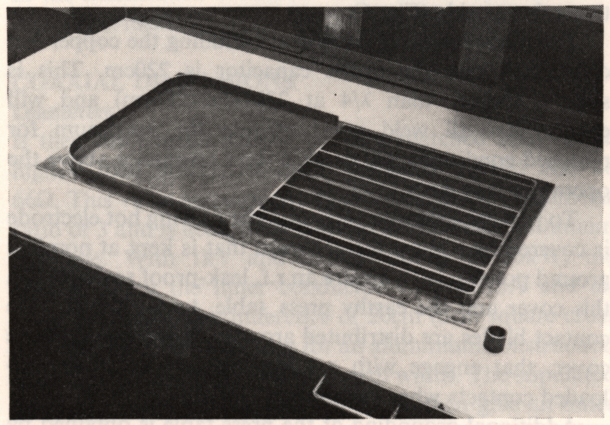

Fig. 6—5cm² electrode (bottom right) and 350cm² electrode on loading tray

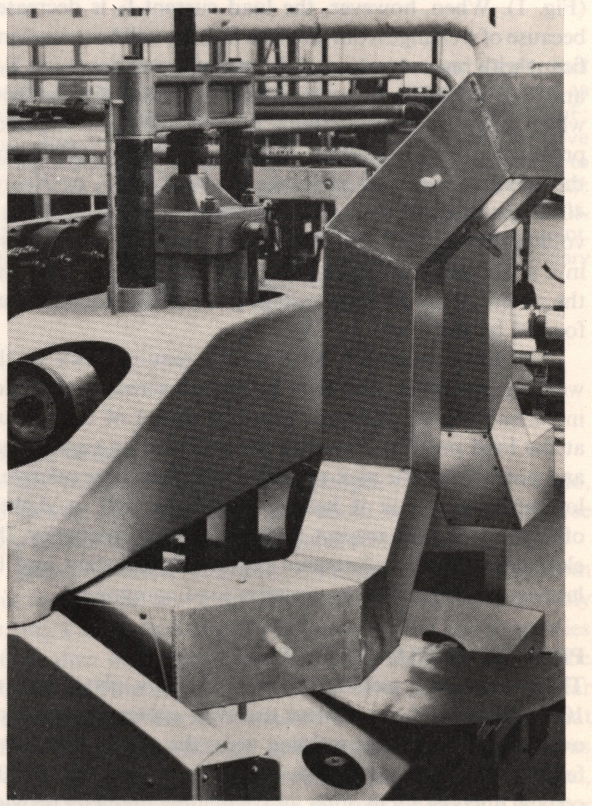

Fig. 7—Press electrode. The bracket in the foreground supports the capacitive divider probe that was used to check the press electrode potential. On the right one set of the flexible electrode connectors may be seen, together with the coaxial inner grounding choke. At the extreme right are two sets of spring-loaded grounding contacts

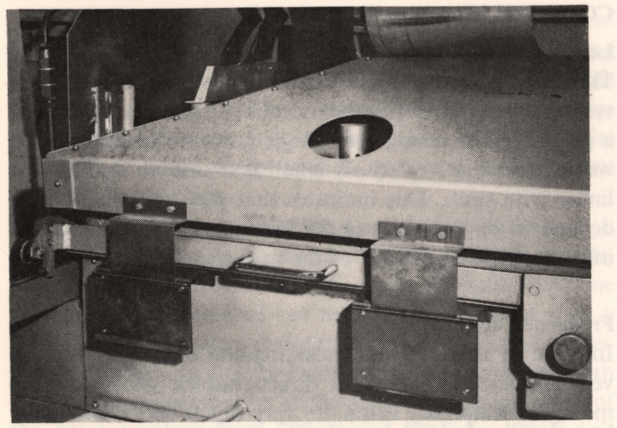

Fig. 8—Close-up of electrode cover grounding contacts in the closed position. The cylindrical pillar protruding from the cut-out can be used as an electrode potential check point

PERFORMANCE

Cold checks

Cold frequency checks were made on the duct resonances for various electrode areas. During these checks the materials under the hot electrode were the same as would be used for a normal weld — two 0·5mm p.v.c. sheets, and one 0·5mm sheet of barrier material.

The resulting duct frequencies are given in Table 2.

TABLE 2
Duct resonances

Weld area cm²	f duct MHz
5	31·4
50	30·6
350	30·05
500	30·0

The duct resonance for all possible welds is thus above the operating frequency of 27MHz.

Operational performance

Practical welds were made in accordance with the design assessment. The materials under the welding electrode were the same as those used for the cold check, namely two layers of 0·5mm p.v.c. and a 0·5mm layer of barrier material.

The probe coupling on the generator end of the coaxial duct was set to give a resultant anode current of 4A at the peak of the weld with the 500cm² electrode. Under these conditions the nominal 6kV anode supply will drop to 5·7kV giving a tube input of 22·8kW. With a d.c. to load efficiency of 70% the peak power during the welding cycle is thus 15·9kW.

For the welding results listed in Table 3 no further circuit adjustments were necessary, except that the timer setting was altered to give optimum results.

TABLE 3
Measurements obtained with three specimen weld areas

Area (cm²) Shape		5 Outline (ring)	350 Outline (car seat)	500 Solid
Time (s)		0·5	4	5
V_a(kV)		6·2	5·8	5·7
I_a(A)	min	0·9	2·5	3
	max	1·1	3·75	4
I_g(A)	max	1·4	1·1	1·0
	min	1·4	0·8	0·7
Electrode potential (kV rms)	max	1·0	1·8	2·1
	min	1·0	1·2	1·1
Δf (kHz)		<50	180	200

Note: The frequency drifts shown above are all contained in a band that extends from 26·5 to 27MHz. (See also frequency spectrum Fig. 9).

All welds were of a good standard with an even weal along the seams between the material layers (Figs. 10 and 11).

The tabulated results appear to be in very good agreement with the original design assessment, especially when it is considered that the dielectric constant as well as the loss factor of the material may change by 100% during the welding cycle.

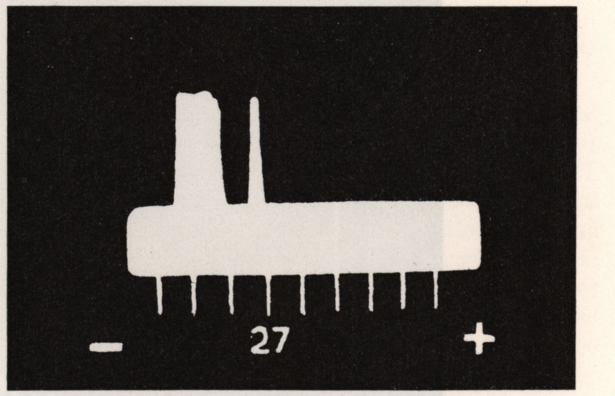

Fig. 9—Frequency spectrum of 5, 350, and 500cm² welds. The traces for the latter two are superimposed. The scale is 200kHz/div., and the welds are within a 500kHz band.

Satisfactory welds with material thicknesses of 2×0·25 and 0·5+0·1mm were also possible. No arcing was experienced under any conditions and no r.f. energy could be detected on the outer skin of the generator or press when a 4662 primed neon probe was used as detector.

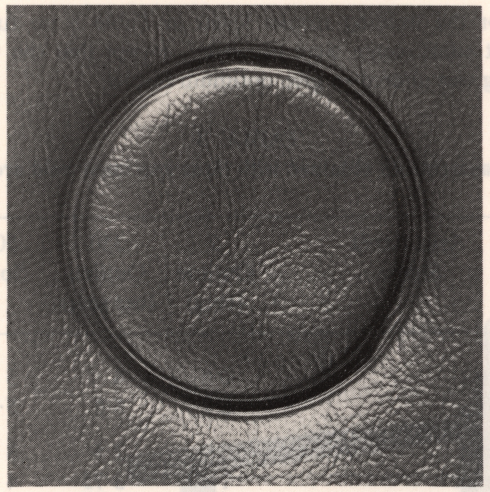

Fig. 10—5cm² weld with electrode potential of 1kV rms, a weld time of 0·5s and a dwell time of 1·0s

COMMENTS

Loading characteristics

The loading characteristics exhibited by this press coupling system appear to justify the use of a $\lambda/2$ generator for such a purpose. The changes of anode current during any one welding cycle are relatively small and smooth and almost linear with time. This indicates that they must be largely dependent on the effect of the increasing temperature that influences the loss factor of the material.

Frequency stability

In spite of the fact that these loading characteristics are very similar to those of a direct-coupled generator, the measured frequency drifts are very much better and show the possibility of containing the drift due to all probable loading conditions well within the approved 27MHz band (Ref. 1).

Tentative trials, during which the transformation ratio of the basic half-wave generator was increased to 1:3, proved this point. With a high voltage end-potential of 18kV, frequency excursions due to all possible loading conditions were kept within a band of 140kHz. Such high potentials are, however, considered unpractical for general industrial use since contamination in an industrial atmosphere could easily lead to arcing and consequent production hold-ups.

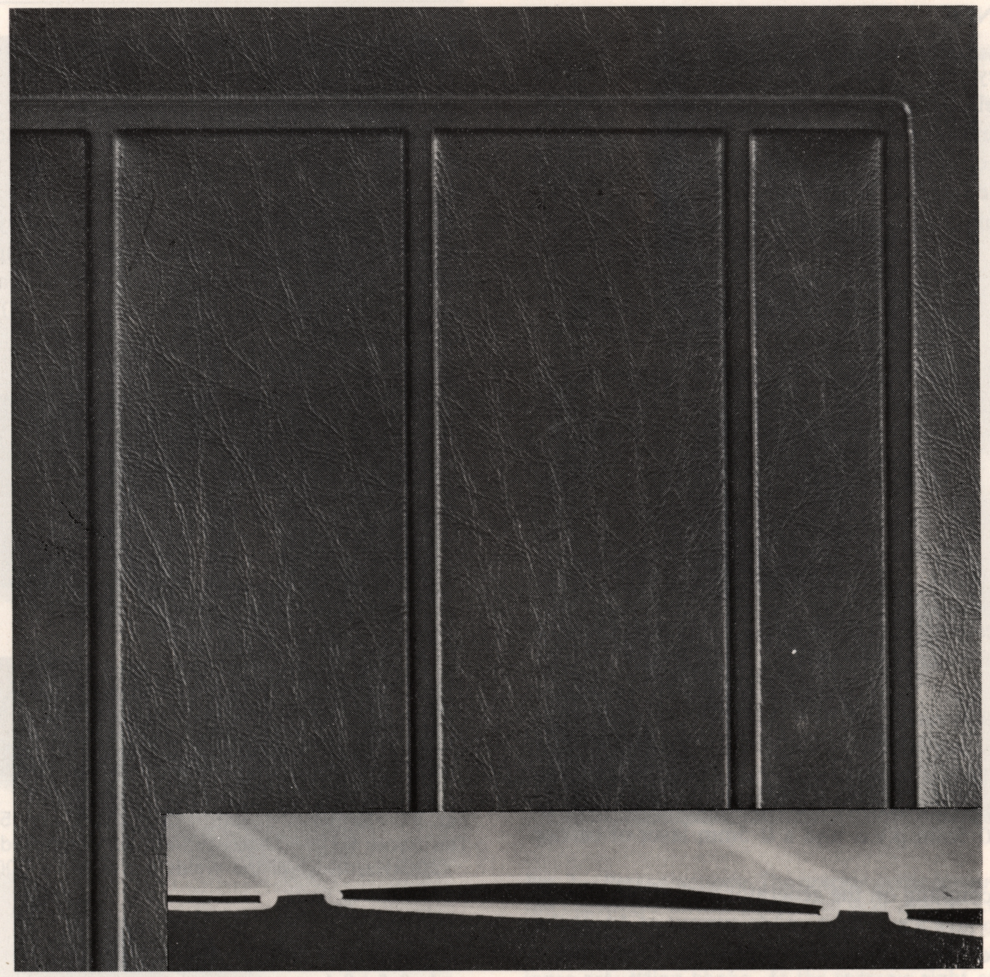

Fig. 11—Part of the 350cm² (car seat cover) weld. The inset shows a section through the weld, viewed at the edge of the sample. The electrode potential was 1·8 to 1·2kV, the weld time was 4s and the dwell time was 3s

There is, however, a safer means of making such a generator conform to the international stability proposals.

In the previous article (Ref. 2) on the half-wave line oscillator it was pointed out that the loaded Q of such a system could, under maintenance of the same transformation ratio (e.g. 1:2), be increased by increasing both end loading capacitances at the same rate. A doubling of the present values from 200pF and 50pF to 400pF and 100pF respectively is certainly feasible. This would double the loaded Q and reduce the above noted frequency shift to something less than 250kHz for all welds. The coupling system between generator and press would not be affected by these modifications because the r.f. potential at the pick-up probe position would remain the same.

Harmonic radiation

Provisional measurements of radiation have shown that the 2nd and 3rd harmonics were at least 50dB and the 4th and 5th more than 40dB below the fundamental.

If further suppression of these or higher harmonics should appear necessary, absorption circuits resonating at the relevant harmonics could be inserted at strategic points in the generator line structure or the coupling duct. Advantage can here be taken of the fact that such harmonic voltage high points may be found that do not coincide with the voltage high points of the fundamental frequency. (See also description of 140MHz mode suppressor in the original $\lambda/2$ generator, Ref. 2 and shown in Fig. 2.)

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1. Dittrich, H. F. 'Tubes for r.f. heating,' Philips, Electronic Components and Materials Division, Eindhoven, 1971, p.26.
2. Dittrich, H. F. 'Dielectric heater using half-wave line at 30MHz' **Mullard Technical Communications,** Vol. 10, No. 94, July 1968, pp. 135 to 145.

Induction heater operating at 2MHz using magnetically beamed triode YD1352S

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UDC 621.365.52.029.53

A 2·5kW induction heater operating at 2MHz has been designed around the magnetically beamed triode YD1352S. The output power can be controlled by a pulse system operating on the gate of the tube between 10% and full output power. The generator is designed with a high Q tank circuit and has no normal anode and grid chokes, and thus gives good operational stability and freedom from parasitic oscillations.

INTRODUCTION

The magnetically beamed triode YD1352S is designed primarily for industrial use to provide a source of radio frequency energy. The tube is mechanically and electrically rugged, can be easily controlled and cannot be damaged by overdrive or overload. These properties have been utilized in the design of a small transportable induction heater operating at 2MHz.

The output control system makes use of the low drive power characteristic of the YD1352S where the grid (which is more accurately known as the 'gate') is pulsed negatively at a variable mark-space ratio. The design also uses 'choke-free' construction techniques which, together with the overload capability of the tube, results in a generator which is virtually foolproof as far as the user is concerned.

YD1352S tube

In the YD1352S the grid is not a grid in the accepted sense but consists of a vertical U shaped flat strip situated so that the parallel sectors are parallel to the cathode. The path of the electrons leaving the cathode is affected by the magnetic field, the electrostatic field due to the anode potential, and the positive grid drive potential (Fig. 1). The combination of electrostatic and electromagnetic control of the electron stream gives a device with a high grid control voltage and a very low grid current—typically 800V drive for a grid current of 2 to 3mA. The d.c. grid biasing arrangement is thus fairly high impedance and is easily varied by a pulse control system. Because the grid is constructed of heavy gauge material and is not a mesh or cage as in conventional tubes, its overload capability is extremely high (twenty times its normal capability at least).

Since the tube requires a very high drive voltage (although the drive power is very low), the anode current of the tube will fall to zero bias value, and the anode dissipation will stay within the dissipation rating, if the feedback voltage in an oscillator circuit should fall to zero because of a circuit fault or a severe overload.

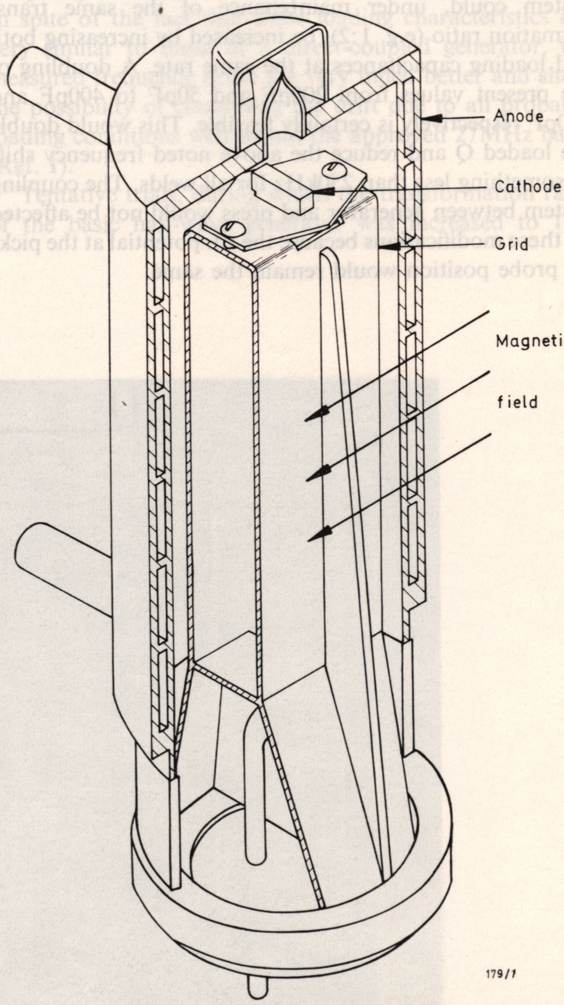

Fig. 1— View of magnetically beamed triode

With: $V_g = 0V$, $p_a = 1{\cdot}25kW$
$V_a = 5kV$, $p_a max = 2{\cdot}0kW$
$I_a = 250mA$,

DESIGN

A value of loaded Q of 100 is chosen for the tank circuit which gives a tank circulating current of approximately 100A. As this is a convenient value for a work coil it is designed to form part of the tank inductance. Because the size of work coil might need to be changed for different loads its inductance is arranged to be not more than 10% of the total tank inductance in order to prevent too great a change in frequency with differing work coils. The circuit has a grounded cathode configuration and the size and layout of the components are arranged for a standard 19in rack-mounted chassis and panel assembly (Fig. 2).

The d.c. supply is full wave rectified and is smoothed by an LC combination. A 100kΩ discharge resistor is connected across the capacitor terminals.

Tank circuit

The suggested operating conditions for the YD1352S using the d.c. smoothed supply are as follows:

$V_a = 5kV$ $R_g = 80k\Omega$
$I_a = 800mA$ $P_{out} = 3{\cdot}1kW$
$I_g = 5mA$

If X_c is the reactance of the tank capacitor then:

$$X_c = (V_a rms)^2/(Q_L \times P_{out}).$$

The rms anode voltage swing is given by:

$$V_a rms = V_a(pk)/\sqrt{2} = 4{\cdot}75/\sqrt{2} = 3{\cdot}36kV$$

where V_a(pk) was measured.

Therefore:

$$X_c = 3360^2/(100 \times 3100) = 36\Omega.$$

Therefore:

$$1/\omega C = 36\Omega$$
$$C = 10^{12}/(2\pi \times 2 \times 10^6 \times 36) = 2{\cdot}22nF.$$

The corresponding inductor L for the tank circuit is given by:

$$\omega L = 36\Omega$$

so: $$L = 2{\cdot}9\mu H.$$

The work-coil inductance is arranged to be 10% of this value, 0·29μH.

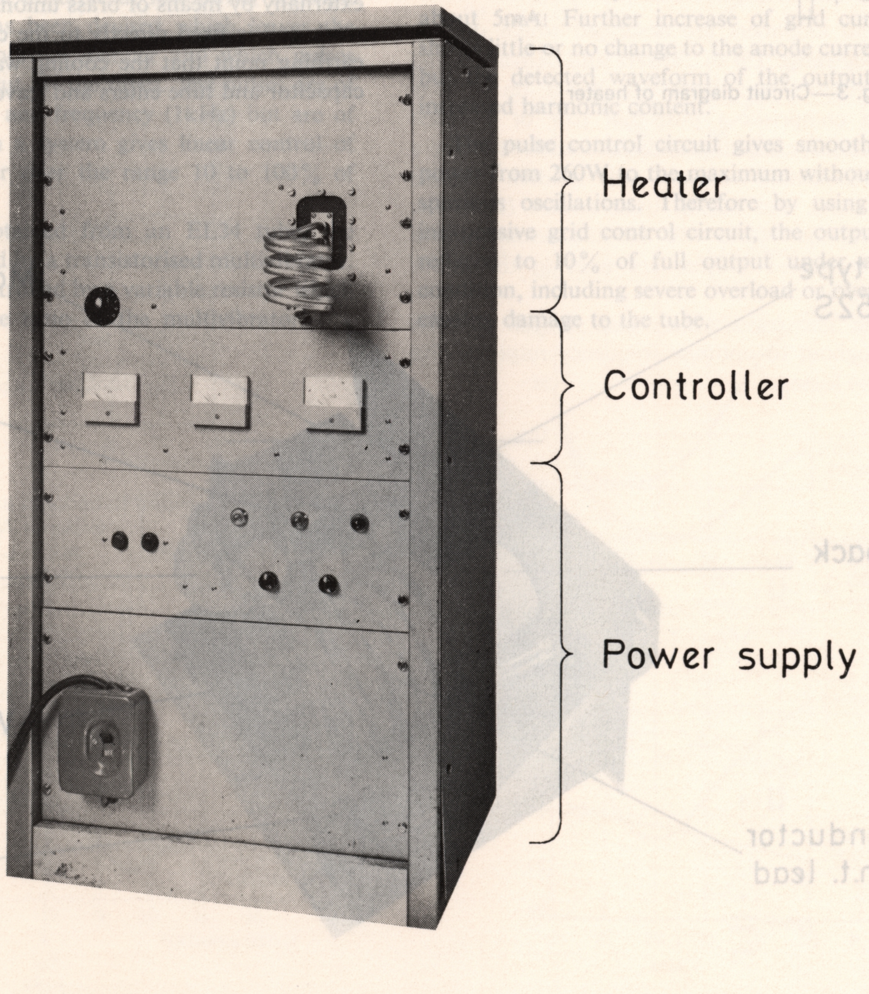

Fig. 2—Front view of heater showing work coil in position

The circuit is given in Fig. 3 and the layout is shown in Fig. 4. The main tank and work coils are constructed from 5/16in (0·8cm) o.d. copper tubing bifilar wound (two tubes in parallel). This provides a go and return circuit for cooling water and also a conductor of reasonable current carrying capacity. The main tank coil is 5in (12·7cm) in diameter and has $5\frac{1}{4}$ turns spaced to give a total length of 5in (12·7cm). The work coil is 2in (6·3cm) in diameter and has 3 turns spaced to give a length of 3in (7·6cm).

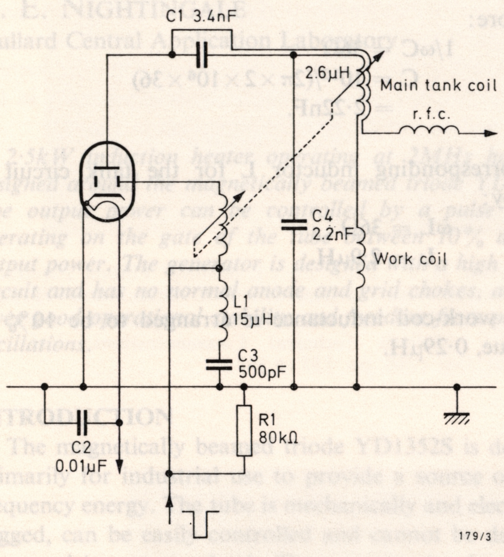

Fig. 3—Circuit diagram of heater

A Valvo water cooled ceramic capacitor type TW85/175, nominal value 2·2nF, is used as the tank capacitor.

Grid circuit

The grid circuit consists of an untuned coil inductively coupled to the tank coil. It is wound on a polypropylene former 3in (7·6cm) in diameter and comprises 10 turns of 16 s.w.g. (1·6mm) copper wire wound at 6 turns per inch. The earthy end of the coil is grounded through a series resonant trap of 500pF and 15µH. This trap allows the grid coil to be grounded to r.f. but offers a high impedance to the grid control pulses mentioned later.

General assembly

Since the tube is operated in this circuit design with a grounded cathode the entire anode block and magnet assembly is at the h.t. potential. The complete assembly, therefore, is mounted on insulating p.t.f.e. pillars. A standard 19in front panel is fixed to the chassis, and the low r.f. potential end of the tank coil is brought to a Tufnol insulator mounted on this panel to enable a work coil, connected in series with the tank coil, to be mounted externally by means of brass unions. The lower end of the work coil is fixed directly to the chassis and it is at this earthing point that the cooling water for the coils, tank capacitor and tube enters and leaves the oscillator.

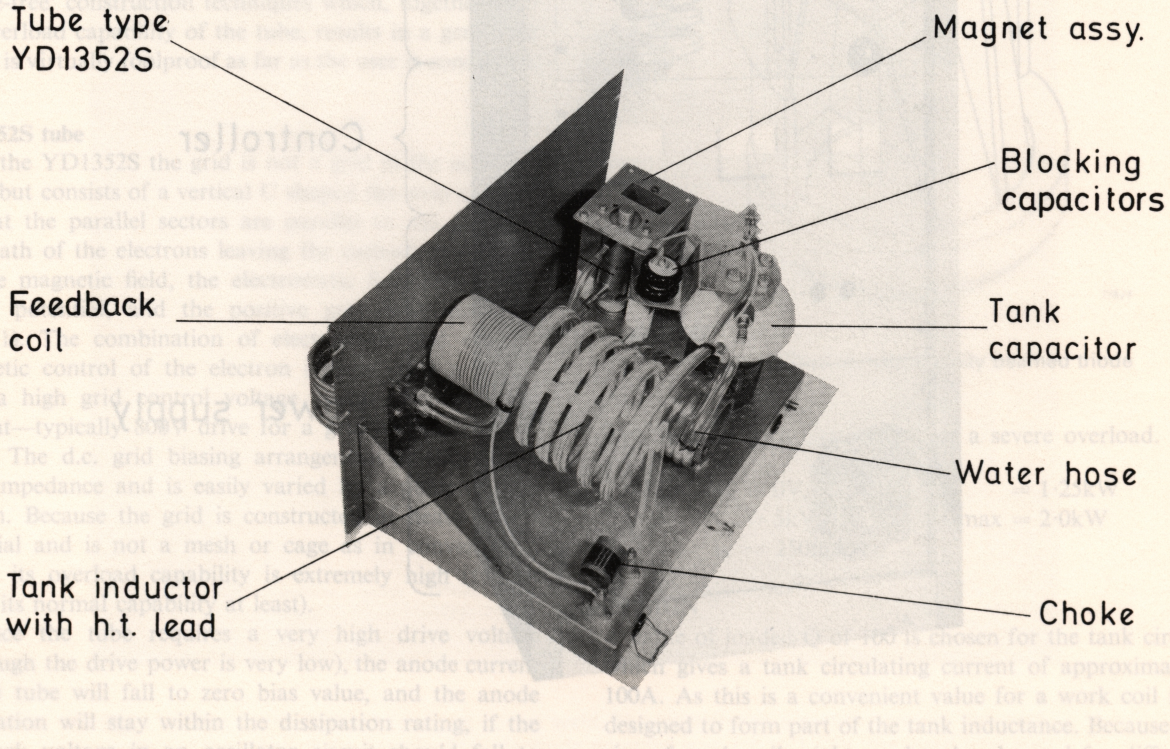

Fig. 4—General layout

The complete tank coil and work coil assembly is kept at d.c. earth potential so that it is necessary to include a blocking capacitor between the live end of the tank coil and the anode. In order that the same cooling water can be used for both the tank and work coils, as well as the tube, a plastic go-and-return hose coil is connected between the live end of the tank coil and the water connections for the tube to give the necessary negligible water leakage path for 5kV d.c. operation. Thus the cooling water flows into the oscillator at earth potential, through the work coil, tank coil and plastic hose coil, to the anode cooler; then back via the hose coil, tank coil and work coil, flowing out at earth potential.

The anode d.c. connecting lead is taken directly across the blocking capacitor (which has a negligible r.f. potential across it) and from there the lead is dressed along the turns of the main tank coil to the junction with the work coil. From this point it is connected to the h.t. input terminal through a small choke. The r.f. potential across the work coil is relatively small so that the choke parameters are not critical.

Because of the low drive power it is possible to control the output power by applying a control potential to the grid. If a simple linear change of bias were used, however, the output would change in a non-linear manner and the oscillator would become unstable at low power settings.

A control circuit has been devised which leaves the tube operating in the normal manner and which pulses the grid negatively (to cut off the tube). The pulses are of fixed amplitude (700V) and frequency (1kHz) but are of variable duration. Such a system gives linear control of the mean output power over the range 10 to 100% of full power.

The pulses were obtained from an EL34 tube, the grid of which is switched by a transistorised multivibrator. The on-off period is controlled by a variable resistor which controls the mark-space ratio of the multivibrator.

By using a thermocouple or similar detecting device on the load a signal can be fed back to the control circuit to maintain a constant temperature in the workpiece.

MEASUREMENTS

To measure the power delivered to an inductively coupled load, a steel can of 2·25in (5·7cm) diameter is arranged to couple by varying degrees to the work coil. A monitored flow of water is passed through the can and an indication of the power dissipated is obtained from the temperature rise of this cooling water.

At the maximum operating conditions for the YD1352S the following result is obtained.

$V_a = 5\text{kV}$ $R_g = 80\text{k}\Omega$
$I_a = 800\text{mA}$ $P_{\text{load}} = 2{\cdot}6\text{kW}$
$I_g = 5\text{mA}$ $\eta_{\text{load}} = 62{\cdot}5\%$

Since the magnetically beamed triode has a grid dissipation rating much higher than is ever likely to be required in practice, the r.f. drive voltage can be varied over a wide range without damage to the tube. Measurements have been made from zero drive to four times the normal grid current. At zero drive (no oscillation) the anode current was 300mA giving an anode dissipation of 1·5kW. As the r.f. drive voltage is progressively increased the oscillator efficiency increases until the grid current is about 5mA. Further increase of grid current to 20mA shows little or no change to the anode current or efficiency but the detected waveform of the output power shows increased harmonic content.

The pulse control circuit gives smooth variations of power from 260W to the maximum without giving rise to spurious oscillations. Therefore by using a simple and inexpensive grid control circuit, the output level can be reduced to 10% of full output under any oscillatory condition, including severe overload or overdrive, without causing damage to the tube.

High-power broadband linear amplifiers using transistors, type 810BLY/A

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UDC 621.375.026.442.029.55

Two broadband linear amplifiers are described using transistors development type 810BLY/A and operating in the frequency range 1·6 to 28MHz. The amplifiers supply output powers of 50 and 100W p.e.p. at power gains of 19 and 16dB respectively. The frequency response of both circuits is flat to within ±1dB, and intermodulation products are better than −27dB at full power.

INTRODUCTION

Two broadband amplifiers are described which operate in the frequency range 1·6 to 28MHz and deliver 50 and 100W p.e.p. respectively. Both are based on a 50W module which comprises a push-pull pair of matched 810BLY/A transistors, a pull-push input transformer and a frequency compensating network. One module together with suitable input and output transformers forms the 50W p.e.p. amplifier while two modules connected in parallel with a hybrid coupler are used in the 100W p.e.p. circuit.

Transmission line transformers and hybrid couplers are used throughout to optimise frequency response and reduce insertion losses.

BASIC 50W MODULE

The basic module, shown in Fig. 1, consists of a matched pair of 810BLY/A transistors operating in class B push-pull, a 9:1 push-pull input transformer and a frequency compensation network to reduce the gain at low frequencies. With values of 10Ω and 2·2nF for R and C, a frequency response flat to within ±1dB over the range 1·6 to 28MHz is obtained.

The 9:1 input transformer provides the best compromise between input match, amplifier gain and frequency response. Design details for this and other transformers are given in a later section.

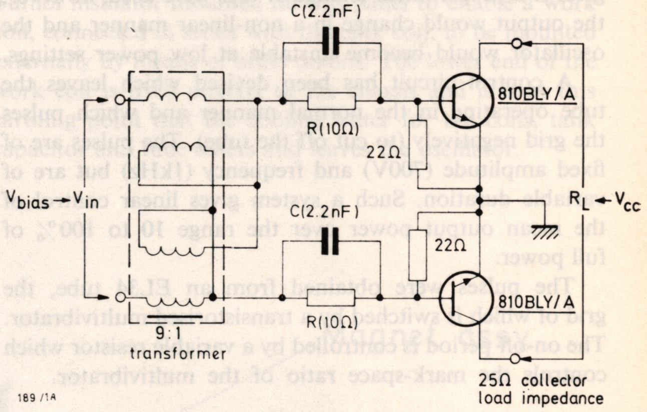


Fig. 1—50W p.e.p. module

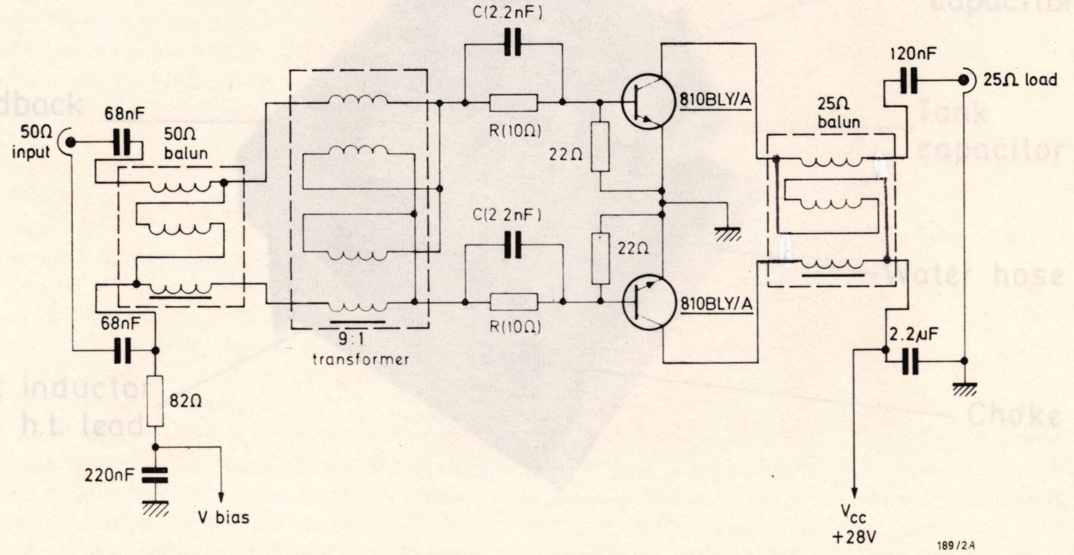


Fig. 2—50W p.e.p. amplifier circuit

50W P.E.P. AMPLIFIER

Design

The addition of a 50Ω balun at the input and a 25Ω balun, together with a 25Ω load at the output, converts the basic module into a 50W p.e.p. amplifier, shown in Fig. 2. The baluns are described in detail on page 67. A bias voltage of about 6V is required and should be derived from the compensated bias supply described later. This bias is adjusted to give an overall amplifier quiescent collector current of 50mA. It is assumed that the collector current is shared equally between the matched transistors.

Careful layout of the amplifier is essential to avoid earth current loops between input and output circuits.

Performance

The performance of the 50W p.e.p. amplifier is summarised in Figs. 3 to 8. A high gain figure of about 19dB is obtained from this circuit with a frequency response level to within ±1dB over the frequency range 2 to 28MHz (Fig. 3). Intermodulation products, measured relative to one tone of a standard two tone test signal, are better than −28dB at 50W p.e.p. output power (Fig. 4). Figs. 5a, b, c and d show distortion plotted against output power at 2, 5, 10 and 27MHz and Figs. 6a and b show the harmonics produced by the amplifier. The efficiency is plotted against frequency and load power in Figs. 7 and 8.

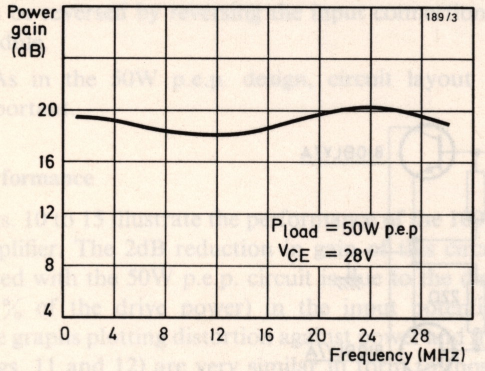


Fig. 3—Power gain as a function of frequency for the 50W amplifier

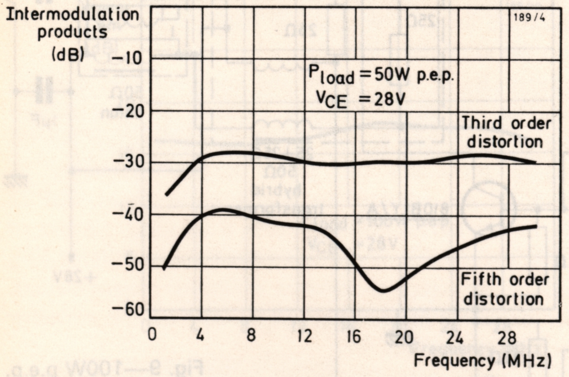
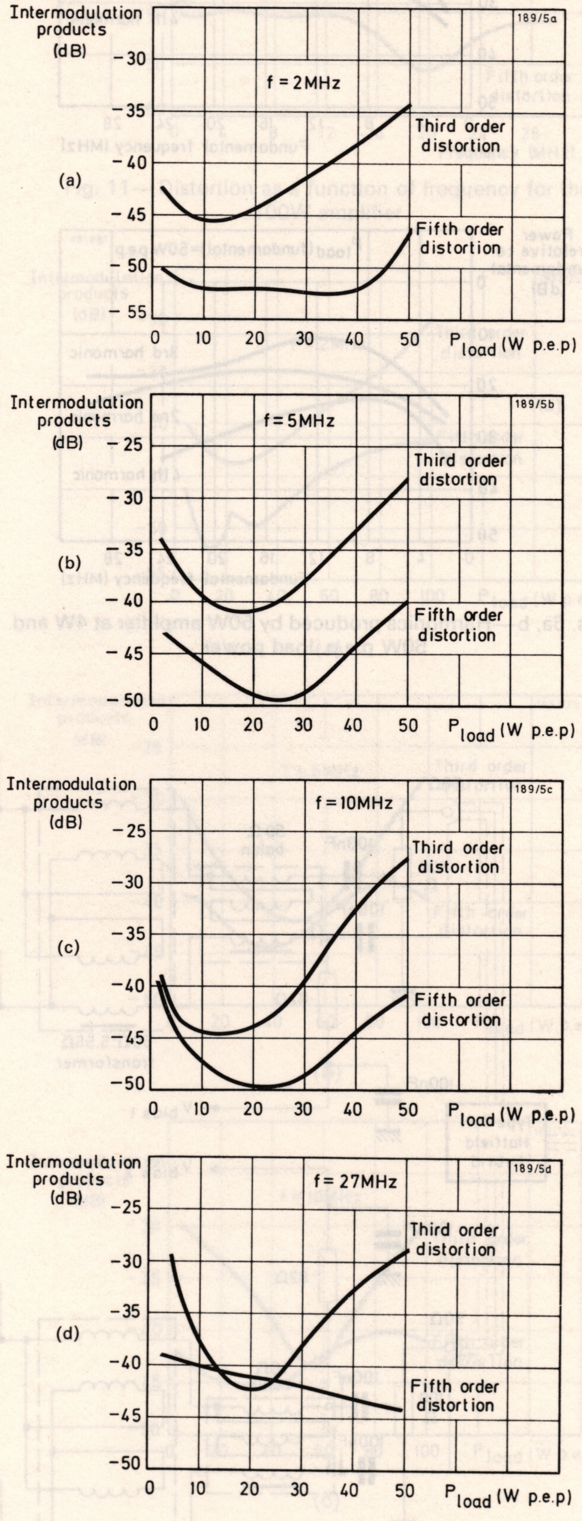
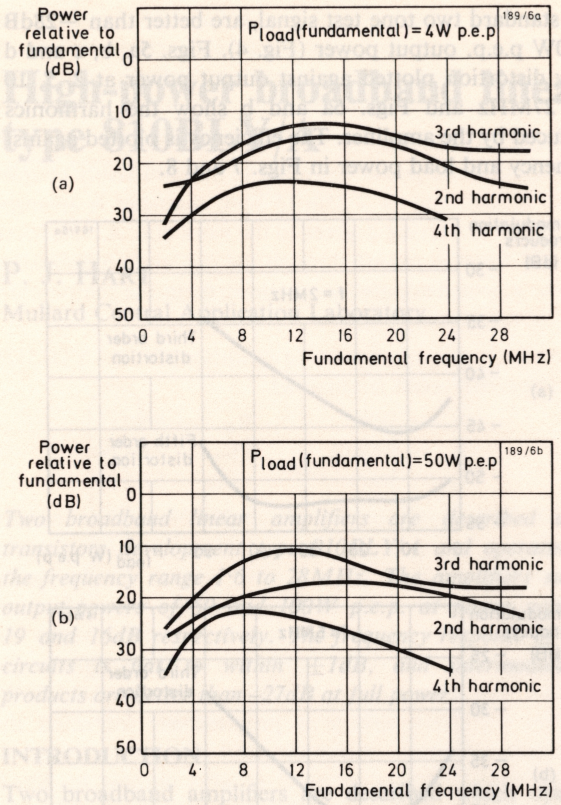


Fig. 4—Distortion as a function of frequency for the 50W amplifier



Figs. 5a, b, c, d—Distortion against output power at 2, 5, 10 and 27MHz for 50W amplifier



Figs. 6a, b—Harmonics produced by 50W amplifier at 4W and 50W p.e.p. load power

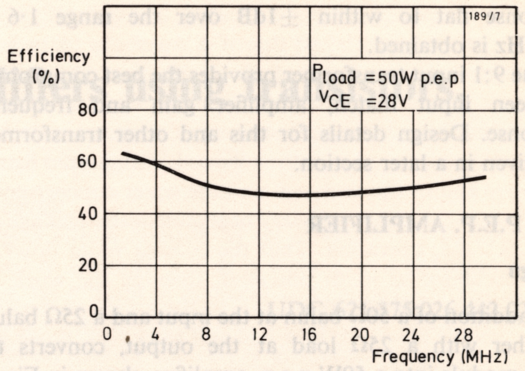


Fig. 7—Two-tone efficiency against frequency for 50W amplifier

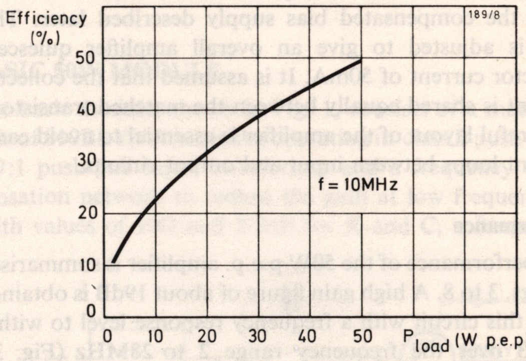


Fig. 8—Two-tone efficiency against load power for 50W amplifier

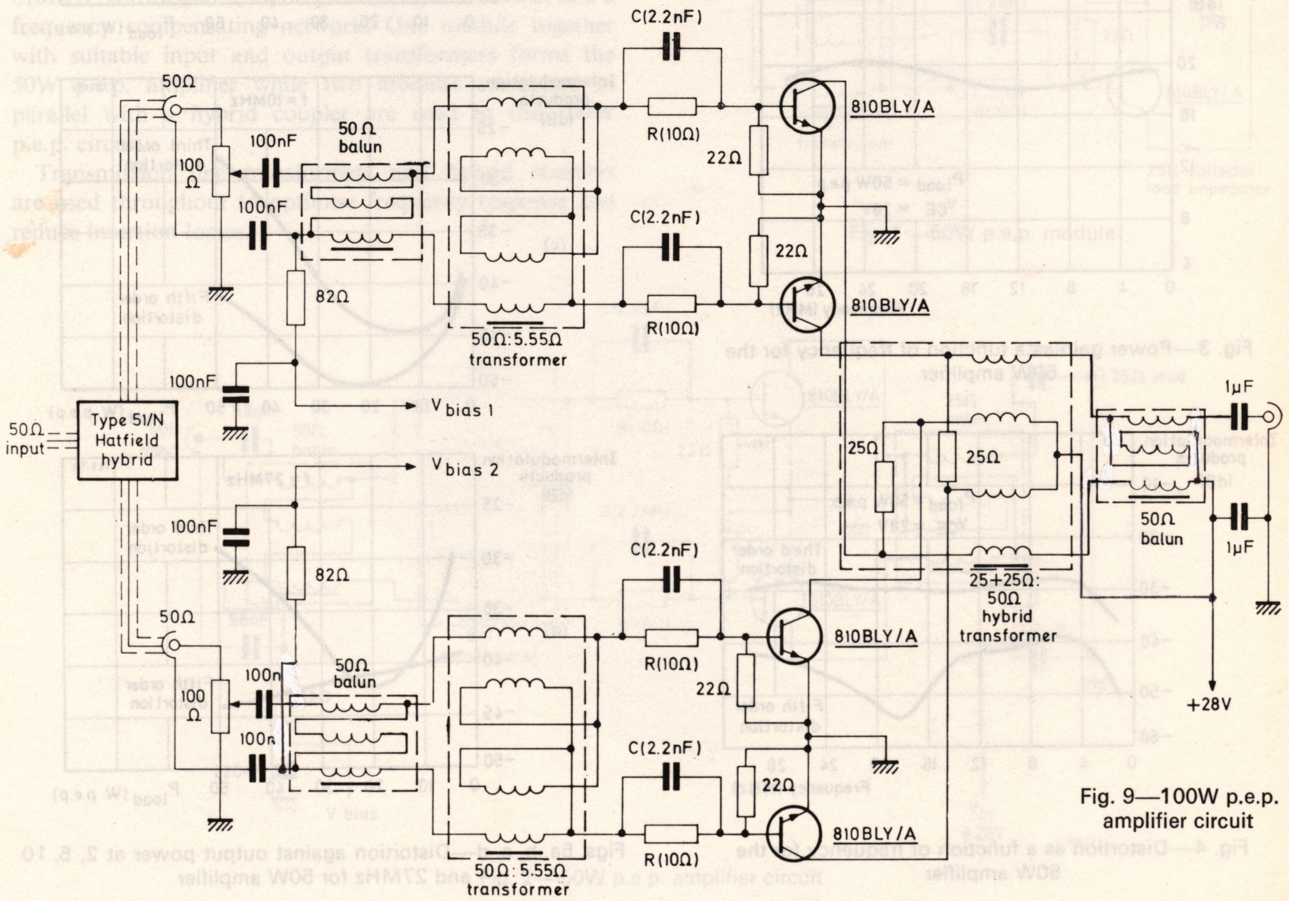


Fig. 9—100W p.e.p. amplifier circuit

100W P.E.P. AMPLIFIER

Design

Two 50W p.e.p. modules are connected in parallel with hybrid couplers to form the 100W p.e.p. amplifier shown in Fig. 9. The output hybrid has an impedance double that of each input. Thus, the output impedance of the amplifier is made to be 50Ω.

The drive signal from a 50Ω source is coupled into a hybrid unit (Hatfield Instrument, type 51/N) which provides two unbalanced outputs isolated at 50Ω. These outputs are connected to 50Ω baluns which provide 50Ω balanced outputs for driving each 50W module. The outputs from the modules are combined in a balanced hybrid coupler and fed through a 50Ω balun to give a 50Ω unbalanced output. Carbon potentiometers in the channel inputs ensure that the gain of both channels is identical. They are adjusted at 10MHz and 100W p.e.p. to give minimum power dissipation in the hybrid resistors together with maximum overall gain.

Separate bias supplies are used for each module, each bias circuit being identical to that described on page 66 and used for the 50W p.e.p. amplifier. The bias supplies are adjusted to give 50mA quiescent current in each module.

Correct operation of the circuit is obtained only if the phasing of the two input signals is correct. The output hybrid transformer also must be wired to give the correct collector current phasing. The phasing in one channel can be reversed by reversing the input connections to one module.

As in the 50W p.e.p. design, circuit layout is again important.

Performance

Figs. 10 to 15 illustrate the performance of the 100W p.e.p. amplifier. The 2dB reduction in gain of this circuit compared with the 50W p.e.p. circuit is due to the dissipation (30% of the drive power) in the input potentiometers. The graphs plotting distortion against power and frequency (Figs. 11 and 12) are very similar in form to those of the 50W p.e.p. circuit.

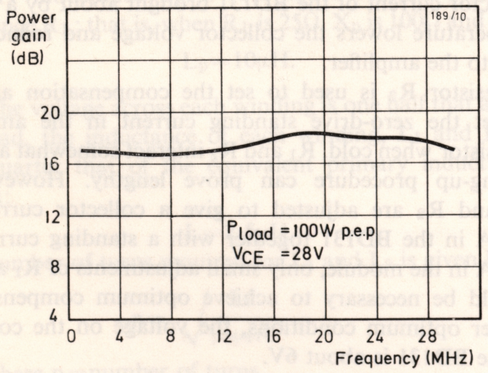


Fig. 10—Power gain as a function of frequency for the 100W amplifier

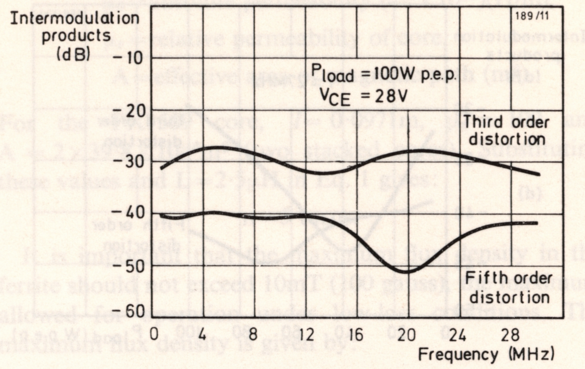
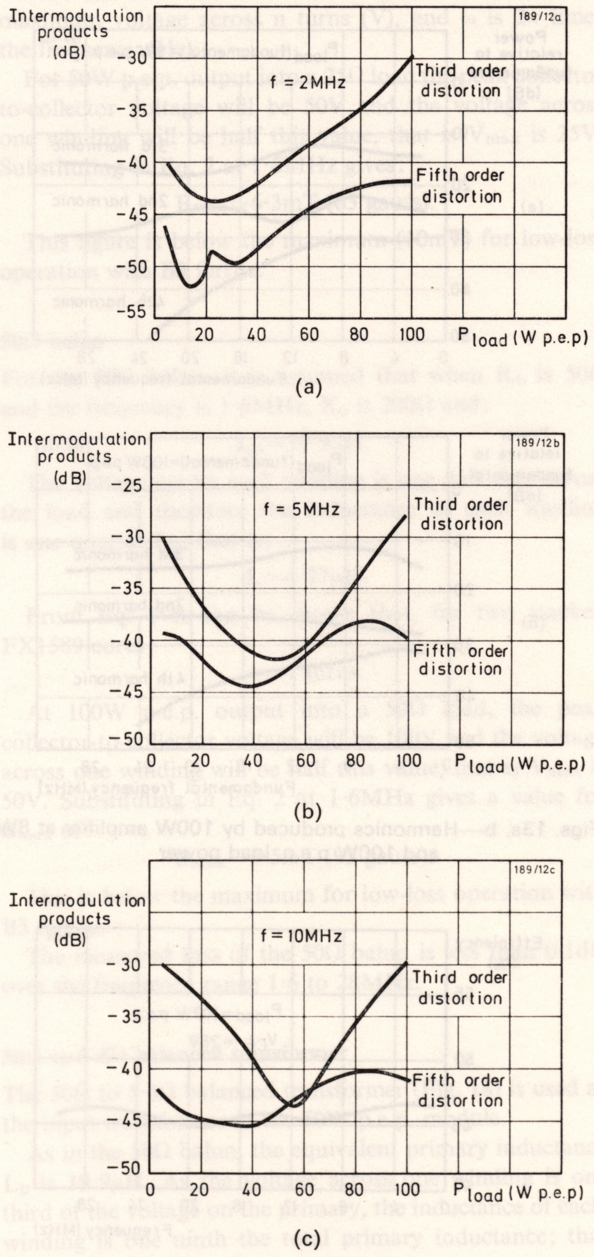


Fig. 11—Distortion as a function of frequency for the 100W amplifier



Figs. 12a, b, c—Distortion against output power at 2, 5 and 10MHz for 100W amplifier

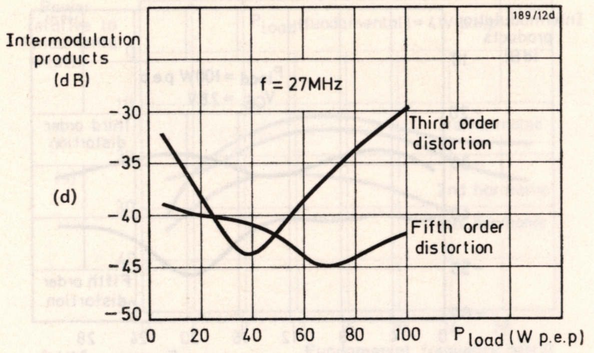
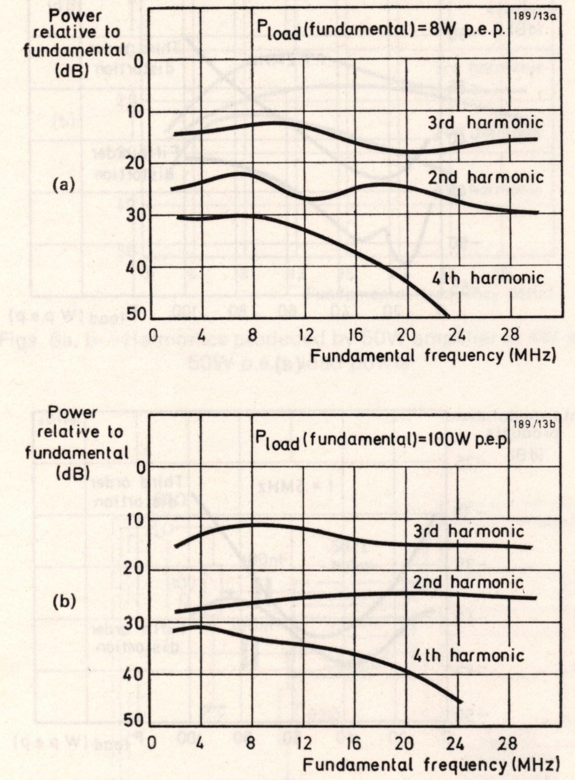


Fig. 12d—Distortion against output power at 27MHz for 100W amplifier



Figs. 13a, b—Harmonics produced by 100W amplifier at 8W and 100W p.e.p. load power

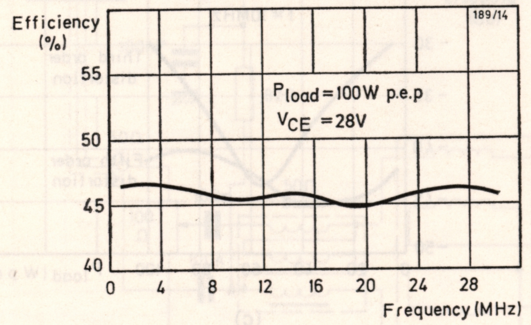


Fig. 14—Two-tone efficiency against frequency for 100W amplifier

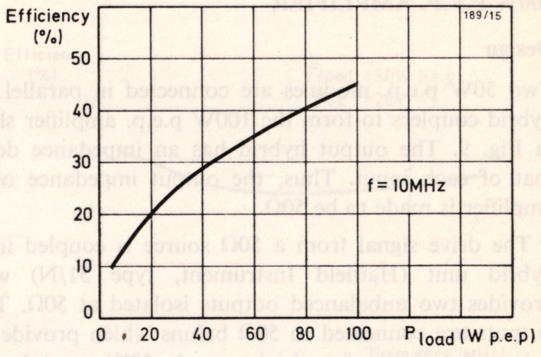


Fig. 15—Two-tone efficiency against load power for 100W amplifier

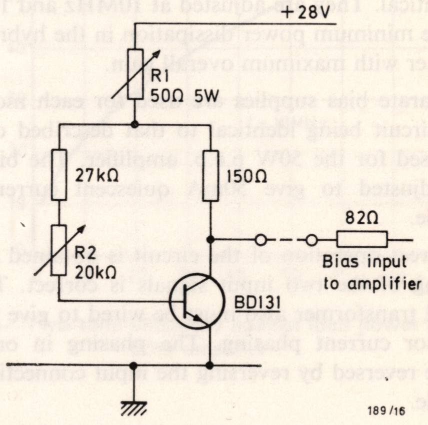


Fig. 16—Bias compensation circuit

BIAS COMPENSATION

Some form of bias compensation is needed to prevent a substantial increase in quiescent collector current when the temperature of the amplifier is raised. A suitable circuit is shown in Fig. 16. The BD131 temperature sensing transistor is mounted on the same heatsink as, and as near as possible to, the 810BLY/A pair it controls. The base current is derived from the 28V supply through a 27kΩ resistor and a 20kΩ variable resistor. An increase in the collector current of the BD131 brought about by a rise in temperature lowers the collector voltage and reduces the bias to the amplifier.

Resistor R_2 is used to set the compensation and R_1 to set the zero-drive standing current in the amplifier transistor when cold. R_1 and R_2 interact somewhat and the setting-up procedure can prove lengthy. However, if R_1 and R_2 are adjusted to give a collector current of 50mA in the BD131 together with a standing current of 50mA in the module, only small adjustments of R_1 and R_2 should be necessary to achieve optimum compensation. Under optimum conditions, the voltage on the collector of the BD131 is about 6V.

It is possible with this circuit to obtain compensation to within ±4mA at 50mA quiescent current over a heatsink temperature range of 16 to 100°C.

TRANSFORMER DESIGN

All the transformers and hybrid couplers used in the amplifiers described are of the transmission line type; that is, the primary and secondary windings are combined to form one or more transmission lines which can be balanced or unbalanced. These transformers offer several advantages over conventional wideband types: very wide frequency response, low insertion losses (typically 0·2dB) and accurate impedance transformation. The main disadvantage is their limited range of transformed impedances, $1/n^2$ compared with n^2/m^2 for conventional transformers (n and m are integers).

The most suitable core material for the wideband transformers used in the amplifiers described is nickel-zinc ferrite, grade B3. This has an effective permeability of between 100 and 200, and low losses at flux densities less than 10mT (100 gauss). 38mm cores, type FX1589, are suitable and these can be stacked to obtain higher power ratings.

Transmission line transformers require the correct impedance cable. The designs to be described use 50Ω, 25Ω and $16\frac{2}{3}\Omega$ cable.

25Ω, 50Ω baluns

The 25Ω balun is used as the output balun in the 50W amplifier. The 50Ω balun is used in the input circuit of both the 50W and 100W circuits and in the output of the 100W amplifier.

In Fig. 17, the windings L_1 and L_2 form the transmission line and winding L_3 is added to provide d.c. continuity.

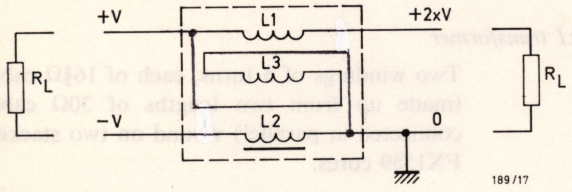


Fig. 17—25, 50Ω balun

It is assumed that at the lowest working frequency of the transformer (1·6MHz), the reactance X_p of the equivalent primary inductance L_p is four times the primary load resistance R_L; that is, when R_L is 25Ω, X_p is 100Ω and:

$$L_p = 10\mu H.$$

As the voltage across each winding is one half that across the load, the inductance of each winding, L_1 and L_2, is one quarter that of the equivalent primary inductance; that is:

$$L = 2{\cdot}5\mu H.$$

The number of turns required for L_1 and L_2 is given by:

$$n = \sqrt{\frac{Ll}{\mu_0 \mu_r A}}, \qquad ..(1)$$

where n = number of turns,

L = winding inductance (H),

l = effective magnetic path length (m),

μ_0 = absolute permeability ($4\pi \times 10^{-7}$)(H/m),

μ_r = relative permeability of core,

A = effective area of magnetic path (m²).

For the FX1589 core, $l = 0{\cdot}0971$m, $\mu_r \simeq 100$ and $A = 2 \times 39{\cdot}8 \times 10^{-6}\text{m}^2$ (two stacked cores). Substituting these values and $L = 2{\cdot}5\mu H$ in Eq. 1 gives:

$$n = 5 \text{ turns.}$$

It is important that the maximum flux density in the ferrite should not exceed 10mT (100 gauss), the maximum allowed for operation under low-loss conditions. The maximum flux density is given by:

$$B_{max} = \frac{V_{max}}{\omega A n}, \qquad ...(2)$$

where B_{max} is the maximum flux density (T), V_{max} is the maximum voltage across n turns (V), and ω is 2π times the frequency (Hz).

For 50W p.e.p. output into a 25Ω load the peak collector-to-collector voltage will be 50V and the voltage across one winding will be half this value, that is V_{max} is 25V. Substituting in Eq. 2 at 1·6MHz gives:

$$B_{max} = 6{\cdot}3\text{mT (63 gauss).}$$

This figure is below the maximum (10mT) for low-loss operation with B3 ferrite.

50Ω balun

For the 50Ω balun, it is assumed that when R_L is 50Ω and the frequency is 1·6MHz, X_p is 200Ω and:

$$L_p = 19{\cdot}9\mu H.$$

The voltage across each winding is one half that across the load and therefore the inductance of each winding is one quarter L_p; that is:

$$L = 4{\cdot}97\mu H.$$

From Eq. 1 it can be shown that, for two stacked FX1589 cores:

$$n = 7 \text{ turns.}$$

At 100W p.e.p. output into a 50Ω load, the peak collector-to-collector voltage will be 100V and the voltage across one winding will be half this value, that is V_{max} is 50V. Substituting in Eq. 2 at 1·6MHz gives a value for B_{max} of:

$$B_{max} = 9{\cdot}5\text{mT (95 gauss).}$$

This is below the maximum for low-loss operation with B3 ferrite.

The measured loss of the 50Ω balun is less than 0·1dB over the frequency range 1·6 to 28MHz.

50Ω to 5·5Ω balanced transformer

The 50Ω to 5·5Ω balanced transformer (Fig. 18) is used as the input transformer in the 50W p.e.p. module.

As in the 50Ω balun, the equivalent primary inductance L_p is 19·9μH. As the voltage across one winding is one third of the voltage on the primary, the inductance of each winding is one ninth the total primary inductance; that is:

$$L = 2{\cdot}21\mu H.$$

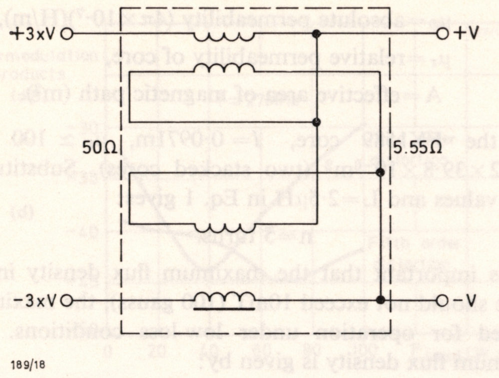


Fig. 18—50 to 5·5Ω balanced transformer

From Eq. 1 it can be shown that, for two stacked FX1589 cores:

$$n=4{\cdot}7 \text{ turns.}$$

B_{max} will be less than 10mT (100 gauss) as in this case the transformer is not required to handle large amounts of power. The required cable impedance is 16⅔Ω and this is approximated by two lengths of 30Ω cable connected in parallel.

The measured loss of the 9:1 transformer is less than 0·25dB over the frequency range 1·6 to 28MHz.

25Ω+25Ω to 50Ω hybrid transformer

The 25Ω+25Ω to 50Ω hybrid transformer (Fig. 19) is used in the output circuit of the 100W p.e.p. amplifier.

It is assumed that the reactance presented across the 50Ω output at 1·6MHz is 200Ω. It can be seen from the voltage distribution in Fig. 19 that the effective primary inductance in parallel with the 50Ω output is 16 times the inductance of one winding. Hence the inductance of one winding is:

$$L=\frac{19{\cdot}9}{16}=1{\cdot}24\mu H.$$

From Eq. 1, at 1·6MHz, it can be shown that, for two stacked FX1589 cores:

$$n=4 \text{ turns.}$$

The maximum power level at each 25Ω input is 50W p.e.p. The maximum voltage across each 25Ω input is therefore 50V and the maximum voltage V_{max} across each winding is 25V. Substituting in Eq. 2 at 1·6MHz gives:

$$B_{max}=7{\cdot}9\text{mT (79 gauss).}$$

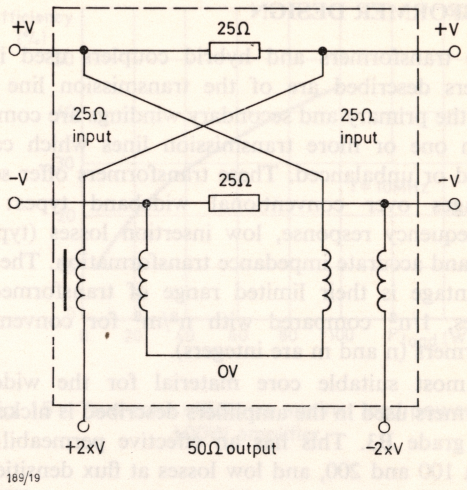


Fig. 19—25+25Ω to 50Ω hybrid transformer

Transformer winding details

25Ω balun: 5 turns of 25Ω cable (Suhner GO1139/25) wound on two stacked FX1589 cores. Third winding consists of 5 turns of 26 swg enamelled copper wire bifilar wound with 25Ω cable.

50Ω balun: 7 turns of 50Ω cable (Suhner GO2232-1, similar to RG174U) wound on two stacked FX1589 cores. Third winding consists of 7 turns of 26swg enammelled copper wire bifilar wound with the 50Ω cable.

9:1 transformer

Two windings of 5 turns, each of 16⅔Ω cable (made up from two lengths of 30Ω cable connected in parallel) wound on two stacked FX1589 cores.

25Ω+25Ω to 50Ω hybrid transformer

Two windings of 4 turns, each of 25Ω cable (Suhner GO1139/25) stacked on two FX1589 cores. The hybrid resistors should each be 25Ω and provided the two 25Ω inputs are identical in both voltage and phase, the resistors dissipate negligible power.